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LINEAR CANCELLATION TECHNIQUE FOR SUPPRESSING IMPULSE NOISE

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I. INTRODUCTION

At the present time, known techniques for suppressing the effects of impulsive disturbances can be identified with one of three general approaches:

(a) The use of nonlinear saturating elements or noise clippers;

(b) Disturbance-triggered, gating-out schemes, whereby the receiver signal path is interrupted at some suitable stage that may fall anywhere from the front end to the input of the back-end transducer; and

(c) The use of special signal design such as special coding, or linear premodulation waveform alteration and postdemodulation correction of the message (for example, pre-emphasis and de-emphasis, presmearing and restoration, and so forth).

Techniques (a) and (b) inevitably leave harmful traces, or "scars", on the message waveform, as well as intermodulation products among the message components and crossmodulation between the desired signal and other signals that may be present. Moreover, these techniques are helpful only when the impulses recur at intervals that are so widely spaced that the disturbance of each impulse dies out before the advent of the next one. Otherwise, the scars in (a) may extend over substantial portions of the message waveform and distort it beyond recognition, or the reception in (b) may be interrupted for extended periods of time.

Technique (c) includes methods having the oft important merit of linearity, and, in principle, can be made sufficiently sophisticated to cope with rapidly recurring impulses - but at the expense of design complexity. Moreover, even in their most sophisticated forms, the predistortion approaches under technique (c) are vulnerable to nonimpulsive disturbances that resemble the smeared forms of the message characters.

In this paper a new technique that is fundamentally simple and linear is described; it is effective even when the i-f filter responses to the various impulses overlap, as long as the number of overlapping responses is not great. The

only important assumption made is the one that is fundamental in all impulse-noise studies -- that the individual bursts have durations that are substantially shorter than the duration of the shortest important message element. This technique (which we call the "linear-cancellation technique") makes use of an important property of the impulse response of a linear filter -- that its maximum value is directly proportional to the filter bandwidth. This technique is applicable to the reception of all types of modulation, even to pulsed transmissions, subject, of course, to the fundamental assumption about the duration of the input noise pulse relative to the duration of the shortest important message element.

II. PRELIMINARY SURVEY

The basis of any effective technique for suppressing extraneous signals is the ability to distinguish between the disturbance and the desired signal and to utilize this distinction automatically. The most important distinguishing property of impulse noise is the fact that the noise energy is initially delivered in very intense bursts of very short duration. For this reason, the fraction-of-time envelope distribution of the noise energy is strongly affected by the frequency characteristics of the medium or filter through which it reaches the observer. The interval of time over which the noise is observed (or the noise power is averaged) has a marked effect upon the percentage-of-time distribution of the noise power.

At the point of observation and comparison with the signal, two important situations may therefore be distinguished; the distinction depends upon whether the impulse noise has been transmitted through a bandwidth that is much wider than the bandwidth of the message elements (as is frequently the case in radio reception), or has unavoidably been processed through a bandwidth that is comparable with the bandwidth of the message elements (as is the case in conventional telephone circuits, for example). If "impulsiveness" is defined to include the attribute of short duration (in comparison with the duration of a message element) as well as high intensity, then in the second-named situation the character of the noise has departed from

impulsiveness, and the unavoidable filtering has reduced the disturbance to an interference from a signal that resembles the desired one.

Techniques that fall under the classifications (a) and (b) of the introduction are effective as long as it is possible to operate on the sum of signal plus impulse noise:

(i) while the duration of the impulsive disturbance is still short compared with the duration of the shortest message element, and

(ii) while the noise peak is appreciably higher than the highest expected signal amplitude.

Both of these conditions are met by the impulse response of a linear wideband transmission filter (see Section III). If either, or both, of these two conditions is seriously violated, the received message may be severely disturbed as a result of too long a "scar" [or too long an interruption], or of no noise clipping [or no gating-out]. In most circumstances in radio communication, both conditions hold at the front end of the receiver, and techniques (a) and (b) have been profitably used. The ultimate performance of each of these approaches can be shown to be a function of the ratio of the preclipping [or pre-gating-out] bandwidth to the postclipping [or postgating-out] bandwidth.

In a typical noise-clipping scheme, the pre-clipping bandwidth may be widened as much as is permissible under the prevailing circumstances, and a clipper with a fixed or a "floating" clipping threshold is arranged to clip off as much of the noise as possible without running into the danger of clipping the signal between the (assumed widely spaced) noise impulses.

In a typical noise-gating-out scheme, the input noise plus signal may be applied at the common input of two receiver signal paths, one of which is essentially a noise-detecting circuit, the other, a circuit whose operation is controlled by the first. Operating under the influence of the incoming impulse noise, the noise-detecting branch gates-out the "controlled" branch completely, as long as the amplitude of the disturbance exceeds a preset threshold. The desired signal emerges from the controlled-branch output with "holes" that may be smoothed out by later filtering.

Under circumstances in which the duration of the impulse-noise disturbance may not differ, to a useful extent, from the duration of a typical message element (as is the case in telephone circuits carrying digital data, for example), approaches (a) and (b) are out of the question. A different approach must be sought, preferably one that anticipates the trouble and designs the signals accordingly. The point of the signal design is to build into the structure of the transmitted waveform an adequate measure of distinguishability from the expected impulse noise. One way to do this is to predistort the signal in

reversible ways, the reverse of which could greatly reduce the impulse-noise disturbance by smoothing it out. This technique is old, and it has been used with remarkable results in the form of pre-emphasis of the message waveform at the transmitter and de-emphasis at the receiver. The need for de-emphasis in order to restore the desired form of the message provides the ability to cut down the postdemodulation (and hence postclipping or postgating-out) bandwidth (if an (a) or a (b) technique is employed) against the noise without, at the same time, hurting the message waveform. In digital-transmission systems, design of signals may involve more "sophisticated" or elaborate operations on the message, such as an error-correction coding operation.

An interesting signal design technique has recently been borrowed from radar.¹ This technique appears promising, especially for use in digital transmission in the presence of narrow-band filtered impulse noise. The basic idea stems from the fact that it is possible to presmear (in time) the message elements at the transmitting end and restore them to their original form at the receiving end by means of complementary smearing filters.¹ The smearing time may be much larger than the duration of a message element. At the receiver, the signal de-smearing (restoring) filter operates on the sum of impulsive disturbance and presmeared signal, and reassembles the signal in its original waveform while it smears out the disturbance. The smearing operation essentially redistributes the energy of an impulse over a "smearing-time" interval that is large compared with the "duration" of the disturbance before smearing and thus reduces the noise intensity in any time interval that equals the duration of a message element. If the impulses recur so rapidly that a large number of them occur during a smearing interval, the statistics of the resulting smeared noise that rides over the restored message elements will approach those of a gaussian-noise process, regardless of the original percentage-of-time distributions of the incoming noise. This suggests a method whereby the nongaussian impulse noise can be transformed into a gaussian noise against which matched-filter detection methods are optimal. The resulting smeared-noise level is greatly abated by standard clipping before the smearing operation. The major drawback of this technique, at the present time, lies in the complexity of the available smearing methods.

III. SOME PROPERTIES OF THE IMPULSE RESPONSE OF BANDPASS FILTERS

We shall review some properties of the impulse response of a linear filter which are of fundamental importance to the present discussion. We shall derive these properties in a very general manner for linear filters of finite bandwidth, and then point out how they can be used to

We start with the general relation

$$h(t) = \frac{1}{2\pi} \int_{-\infty}^{\infty} H(j\omega) e^{j\omega t} d\omega \quad (1)$$

which relates the impulse response $h(t)$ of a linear filter to the system function $H(j\omega)$ of the filter. Using the subscripts BP and LP to denote bandpass and lowpass, we write

$$H_{BP}(j\omega) = \begin{cases} H_{LP}(j\omega - j\omega_0), & \omega_1 \leq |\omega| \leq \omega_2, \\ \omega_0 = (\omega_2 - \omega_1)/2 \\ 0, & \text{elsewhere} \end{cases} \quad (2)$$

Then

$$\begin{aligned} h_{BP}(t) &= \frac{1}{2\pi} \left[\int_{-\omega_2}^{-\omega_1} H_{LP}(j\omega + j\omega_0) e^{j\omega t} d\omega \right. \\ &\quad \left. + \int_{\omega_1}^{\omega_2} H_{LP}(j\omega - j\omega_0) e^{j\omega t} d\omega \right] \\ &= \frac{1}{\pi} \int_{\omega_1}^{\omega_2} \operatorname{Re} \left\{ H_{LP}(j\omega - j\omega_0) e^{j\omega t} \right\} d\omega \end{aligned}$$

Now, if we let $x = (\omega - \omega_0)/(\omega_2 - \omega_1)/2$, then x measures the frequency deviation $\omega - \omega_0$ from the center frequency ω_0 in units of one-half the bandwidth $(BW) = \omega_2 - \omega_1$, $\omega = \omega_1$ goes to $x = -1$, $\omega = \omega_2$ goes to $x = 1$, and if we set $H_{LP}(jx(BW)/2) = A_{LP}(x)e^{-j\phi_{LP}(x)}$, we can write

$$\begin{aligned} h_{BP}(t) &= \frac{(BW)}{2\pi} \int_{-1}^1 A_{LP}(x) \cos \left[x \frac{(BW)}{2} t - \phi_{LP}(x) + \omega_0 t \right] dx \\ &= a_c(t) \cos \omega_0 t - a_s(t) \sin \omega_0 t \\ &= \sqrt{a_c^2(t) + a_s^2(t)} \cos \left[\omega_0 t - \tan^{-1} \frac{a_s(t)}{a_c(t)} \right] \end{aligned} \quad (3)$$

where

$$a_c(t) = \frac{(BW)}{2\pi} \int_{-1}^1 A_{LP}(x) \cos \left[x \frac{(BW)}{2} t - \phi_{LP}(x) \right] dx \quad (4)$$

and

$$a_s(t) = \frac{(BW)}{2\pi} \int_{-1}^1 A_{LP}(x) \sin \left[x \frac{(BW)}{2} t - \phi_{LP}(x) \right] dx$$

For the special case in which $A_{LP}(x)$ is an even function of x and $\phi_{LP}(x)$ is an odd function, $a_s(t) = 0$, and Eq. (3) reduces to

$$h_{BP}(t) = 2 h_{LP}(t) \cos \omega_0 t \quad (5)$$

in which we have set $a_c(t) = 2h_{LP}(t)$.

Let us now examine the preceding expressions. The restriction of the system function of the filter as stated in Eq. (2) is a tolerable idealization for practical receiver filters. The purpose of Eq. (2) is to emphasize the frequency-selective character of the filter response; therefore ω_2 and ω_1 are to be interpreted as limits outside of which the filter amplitude-response characteristic is negligible for all practical purposes. The frequency ω_0 is, by definition, the midband frequency, and according to Eqs. (3) and (5) it has the significance of a carrier frequency for the impulse response of the filter. The assumption of even and odd symmetry for $A_{LP}(x)$ and $\phi_{LP}(x)$ is realistic when the natural frequencies of the filter are symmetrically disposed with respect to the line $s = j\omega_0$ and $\omega_2 - \omega_1 \ll \omega_0$ (the so-called high-Q approximation). When this assumption can be made without serious error, as is most often the case, the impulse response of the filter can be interpreted as being a constant-frequency amplitude-modulated sinusoid whose instantaneous amplitude is given by two times the impulse response of a lowpass filter whose system function is

$$H_{LP}(j\omega) = \int_{-\infty}^{\infty} h_{LP}(t) e^{-j\omega t} dt \quad (6)$$

The most important properties of the instantaneous amplitude of the impulse response can be seen from Eqs. (4) and (5). According to the expression in Eq. (4), the time variable t in any final expression for $h_{LP}(t)$ will always be multiplied by a scaling factor that equals some constant times any convenient measure of filter bandwidth (be it the separation between the half-power points, or the separation between the frequencies ω_1 and ω_2 which were described above). One consequence of this time scaling

is that if the filter bandwidth is scaled up by some factor, the time scale of the impulse response will be scaled down by the same factor. Also, since $A_{LP}(x)$ is positive for all x , the absolute maximum of $h_{LP}(t)$ occurs when the cosine factor attains its first maximum value of unity; that is, when

$$t = t_m = \frac{\phi_{LP}(x)}{x} / (BW)/2 \quad (7)$$

Accordingly, the maximum value of $h_{LP}(t)$ is

$$h_{LP}(t_m) = \frac{(BW)/2}{2\pi} \int_{-1}^1 A_{LP}(x) dx \quad (8)$$

Equations (7) and (8) show that the maximum value of the envelope of the impulse response is directly proportional to the filter bandwidth, and the instant of time at which the maximum occurs is inversely proportional to the filter bandwidth. This statement holds independently of whether or not $A_{LP}(x)$ and $\phi_{LP}(x)$ have the appropriate symmetries about $x = 0$, and the stated types of proportionality hold for any convenient choice of bandwidth-measurement criterion.

Finally, we note from Eq. (6) that

$$\int_{-\infty}^{\infty} h_{LP}(t) dt = H_{LP}(0)$$

which shows that the area under the instantaneous amplitude of the impulse response is a constant that is fixed by the center-frequency value of the filter amplitude response.

The preceding results apply directly to the impulse-noise problem. If the impulsive input is of such a short duration that its spectrum is substantially flat over the range $\omega_1 \leq \omega \leq \omega_2$, then the response of the linear filter will differ from $h(t)$ only by a constant multiplier. We can therefore draw the following conclusions:

(a) The impulsive disturbance causes a response that is generally a combined amplitude and frequency-modulated sinusoid with a carrier frequency that equals the center frequency of the filter passband. If the filter amplitude and phase characteristics are closely approximated by even and odd functions, respectively, of frequency deviation from the center frequency, then the frequency modulation is negligible, and the envelope of the response is given by two times the impulse response of a physical lowpass analog filter.

(b) The maximum value of this envelope is directly proportional to the filter bandwidth, and the duration of its important part is inversely proportional to this bandwidth--the criteria of bandwidth and duration measurements being dic-

tated entirely by arbitrary considerations related to the practical situation at hand. Thus, if the nominal bandwidth of the filter is multiplied by a factor k , the maximum value of the impulse response is multiplied by k , and the nominal duration of this response by $1/k$. The area under the impulse response is a constant independent of this scaling.

IV. THE BASIC LINEAR IMPULSE NOISE SUPPRESSION TECHNIQUE

These results can be applied directly, as illustrated in Fig. 1, to suppress impulse noise. The indicated operation can be performed either in the r-f section or in the i-f section of a receiver, even though for practical reasons, as well as for reasons of receiver sensitivity to fluctuation noise, the i-f section may be preferable. The design of each of the paralleled amplifiers is assumed to satisfy the linearity and bandwidth requirements for low-distortion transmission of the desired signal $e_s(t)$. If the minimum bandwidth required to accommodate the signal $e_s(t)$ is B cps, then one amplifier is designed to have this bandwidth and a gain of G , the other is designed to have a bandwidth kB , $k > 1$, and gain bG . Direct subtraction of the amplifier responses leads to an output-signal component given by $G(1 - b)e(t)$ and an impulsive-noise component given by $N_o(t) = N_1(t) - N_2(t)$. The maxima of $N_1(t)$ and $N_2(t)$ will attain equal values if $b = 1/k$, but they will occur at instants whose spacing depends on k . An appropriate choice of b should therefore enable the subtraction to result in an $N_o(t)$ that is substantially milder than $N_1(t)$ when compared with the desired signal. Bandwidth restriction to the minimum necessary value of B in circuits that follow the arrangement shown in Fig. 1 is expected to result in the removal of any added random noise or interference that may pass through the excess bandwidth of the lower amplifier branch. The circuits that precede the arrangement of Fig. 1 should preferably have a bandwidth that is considerably wider than kB , in order to preserve the impulsive character of the input disturbance until it reaches the input terminals of the paralleled amplifiers.

V. ANALYSIS OF SOME SPECIFIC CIRCUITS

We shall now examine the effectiveness of some specific circuit embodiments of the new technique. The ultimate criterion for judging the effectiveness of specific circuits is, of course, some measure of the reduction of the noise effects upon the receiver output. For example, with an FM receiver the ultimate performance criterion may be the percentage of reduction in the capture time (or the time interval during which the response of the filter to the impulsive disturbance exceeds the signal amplitude) because the probability that the disturbance will cause a "pop" noise in the output is directly

proportional to the capture time. It will be found, however, that the improvement in the ratio of signal amplitude to the peak value of the noise amplitude as a result of the subtraction is a much easier quantity to compute, as well as a criterion that is more generally applicable to systems other than FM.

Consider the situation in which each of the parallel branches in Fig. 1 is made up of an amplifier with a high-Q single-tuned circuit as the load. The instantaneous amplitude of the unit-impulse response of such an amplifier can be written as

$$2 h_{LP}(t) = 2 a G e^{-at} \quad (9)$$

where a is one-half the bandwidth of the circuit measured between the half-power points, and G is the gain of the amplifier at its center frequency. Thus, if the duration of the input disturbance $N_1(t)$ in Fig. 1 is so short relative to the time constants of the paralleled amplifiers that the spectrum of $N_1(t)$ is essentially flat over the significant passband of the lower-Q circuit, and if the magnitude of the Fourier transform of $N_1(t)$ within this passband is A_N , then $N_1(t)$ simulates an impulse of strength A_N , and the instantaneous amplitudes of $N_1(t)$ and $N_2(t)$ are $2 a A_N G e^{-at}$ and

$2 a A_N b k G e^{-kat}$, respectively. Since both of these responses attain their maximum values at the same instant of time, we choose $b = 1/k$ for a perfect cancellation at $t = 0$.

The instantaneous amplitude of $N_o(t)$ with $b = 1/k$ is

$$E_N(t) = 2 a A_N G \left[e^{-at} - e^{-kat} \right] \quad (10)$$

The function

$$f_1(x) = e^{-x} - e^{-kx} \quad (11)$$

is plotted in Fig. 2 for various choices of k . The peak value of $f_1(x)$ occurs when

$$x = x_m = \ln k^{1/(k-1)}$$

and

$$f_1(x) \Big|_{\max} = (k-1)k^{-k/(k-1)} \quad (12)$$

Therefore, the maximum value of the output noise envelope, $E_N(t)$, is

$$E_N(t) \Big|_{\max} = 2 a A_N G (k-1)k^{-k/(k-1)} \quad (13)$$

In contrast, if the input signal is a sinusoid of amplitude E_s , the output signal amplitude is $G E_s (k-1)/k$. The ratio of output signal amplitude to peak noise amplitude is then

$$\left[\frac{S_{\text{peak}}}{N_{\text{peak}}} \right]_{\text{out}} = \left[\frac{E_s}{2 a A_N} \right] k^{1/(k-1)} \quad (14)$$

The quantity in brackets on the right is recognized as the ratio of signal amplitude to peak noise at the output of the upper branch in Fig. 1.

The factor $k^{1/(k-1)}$ represents the improvement achieved through subtracting the output of the lower amplifier from the output of the upper amplifier. This improvement factor is a monotonically decreasing function of k for $k > 1$. The maximum value, which is approached as $k \rightarrow 1$ from above, can be found by setting $k = 1 + \epsilon$, and letting $\epsilon \rightarrow 0$. This results in

$$\lim_{\epsilon \rightarrow 0} (1 + \epsilon)^{1/\epsilon} = e = 2.718 \dots$$

= base of natural logarithms

We therefore conclude that since k is, by hypothesis, always chosen greater than one ($k = 1$ results in complete cancellation of signal as well as disturbance), the improvement factor will be less than 8.686 db. For $k = 1.2$, it is 8 db, and for $k = 1.1$ it is 8.28 db.

Note that it is possible, by appropriate pre-clipping of the incoming noise peaks, to keep $(E_s/2 a A_N)$ sufficiently low that the ratio of peak signal to peak noise (Eq. 14) after the cancellation in Fig. 1 always exceeds unity by a comfortable margin. This would effectively eliminate all pops in FM and all confusion (in gates with preset thresholds) between impulsive noise pulses and the desired pulses in pulse transmission, and to improve other types of reception well beyond the limits on improvement with noise clipping alone.

As a second example, we consider the situation in which each of the parallel branches in Fig. 1 is made up of n synchronously tuned stages in cascade, each having a single-tuned high-Q circuit with a damping factor a for the upper branch, and ka for the lower branch. Let the overall gains be G_n for the upper branch, and G_n/k for the lower branch. If we take the expression in Eq. (9) for the envelope of the impulse response of the first stage in the upper branch and convolve it with $h_{LP}(t)$ of the second stage, the outcome will be the envelope of the impulse response of the first two stages in cascade. Repetition of this procedure leads to

$$a_{nu}(t) = \frac{2 a G_n A_N}{(n-1)!} (at)^{n-1} e^{-at} \quad (15)$$

for the envelope of the over-all response of the upper branch. The peak of this envelope is

$$2\alpha GA_N \frac{1}{(n-1)!} \left(\frac{n-1}{e}\right)^{n-1} \quad (16)$$

and it occurs at $t = (n-1)/\alpha$. Thus, the ratio of signal amplitude to peak-noise amplitude at the output of the upper branch is

$$\begin{aligned} \rho_n &= \frac{E_s}{2\alpha A_N} \left(\frac{e}{n-1}\right)^{n-1} (n-1)! \\ &= \frac{E_s}{(BW)A_N} \psi(n) \end{aligned} \quad (17)$$

where (BW) is the over-all bandwidth in rad/sec, and

$$\psi(n) = (n-1)! \left(\frac{e}{n-1}\right)^{n-1} \sqrt{2^{1/n} - 1} \quad (18)$$

The function $\psi(n)$ represents an improvement factor (relative to the output of one stage of bandwidth (BW)) that is characteristic of cascading synchronous, single-tuned, high-Q amplifiers. Some values of $\psi(n)$ are: $\psi(1) = 1$, $\psi(2) = 1.75$, $\psi(3) = 1.88$, $\psi(n) \approx 2.1\sqrt{1-(1/n)}$ for $n > 3$.

The envelope of the impulse response of the scheme illustrated in Fig. 1, in the present example, is therefore

$$\begin{aligned} E_N(t) &= \frac{2\alpha GA_N}{(n-1)!} \left[(at)^{n-1} e^{-at} \right. \\ &\quad \left. - (kat)^{n-1} e^{-kat} \right] \end{aligned} \quad (19)$$

The determination of the peak value of $E_N(t)$ is not a simple task here. But our major interest at this point is only in the upper bound on the improvement (that is attributable to the subtraction) in the ratio of signal amplitude to peak value of $E_N(t)$. This upper bound is the limit as $k \rightarrow 1$ from above, of the measure of improvement for any $k > 1$. The desired upper bound is therefore most expeditiously obtained by setting $k = 1 + \epsilon$ in Eq. (19) and then operating on an approximation to $E_N(t)$ that holds for $0 < \epsilon \ll 1$. If we set

$$f_n(x) = x^{n-1} e^{-at} - (kx)^{n-1} e^{-kx}, \quad (20)$$

then the desired approximation for $E_N(t)$ is

$$\begin{aligned} E_N(t) &\approx \frac{2\alpha GA_N}{(n-1)!} \epsilon \\ &\quad \text{times} \left\{ x^{n-1} e^{-x} [x - (n-1)] \right\}. \end{aligned} \quad (21)$$

Let the function in braces be denoted $g(x)$. The upper bound on the value of the ratio of signal amplitude to peak-noise amplitude after subtraction is then

$$\frac{\frac{E_s}{2\alpha A_N}}{(n-1)!} \left/ \frac{[g(x)]_{\max}}{(n-1)!} \right. \quad (22)$$

In contrast, the ratio of signal amplitude to peak noise amplitude at the output of the upper branch is

$$\frac{\frac{E_s}{2\alpha A_N}}{(n-1)!} \left/ \left(\frac{n-1}{e}\right)^{n-1} \right. \quad (23)$$

The improvement factor attributable to subtraction is therefore bounded by $|I_n|$, where

$$\begin{aligned} I_n &= \left(\frac{n-1}{e}\right)^{n-1} \left/ [g(x)]_{\max} \right. \\ &= \frac{4(n-1)^{n-1} e^{\frac{1}{2}[1 \pm \sqrt{4n-3}]} }{6n-4 \pm 2n\sqrt{4n-3}} \end{aligned} \quad (24)$$

in which the sign that yields the smaller of the two values of $|I_n|$ is chosen (minus for $n \geq 3$).

Illustrative values of $|I_n|$ are: $I_1 = e$, which

checks the result of the first example (Eq. (14) et seq.), $I_2 = 1.2$, $|I_3| = 4/e = 1.47$, $|I_4| = 3.33$,

$|I_5| = 14.2$, $|I_7| \approx 790$. The improvement builds up very rapidly with increasing n for $n > 3$.

We now illustrate the effect of timing the peak values of the envelopes of the impulse responses of the paralleled amplifiers so that they occur at the same instant of time. We take the case in which $n = 2$ in the preceding example. Here, the instantaneous amplitude of $N_o(t)$ depends upon at according to the function

$$f_2(x) = xe^{-x} - kxe^{-kx} \quad (25)$$

which is plotted in Fig. 3(a) for various values of k . The peak values of xe^{-x} and kxe^{-kx} are attained at $x = 1$ and $x = 1/k$, respectively. Through the use of appropriate delay, kxe^{-kx} can be made to attain its peak value at $x = 1$, in order to null the peak value of xe^{-x} . If such delay is introduced into the lower branch then $f_2(x)$ is replaced by

$$f_{2d}(x) = xe^{-x} - [kx - (k-1)]e^{-[kx-(k-1)]} \quad (26)$$

which is plotted in Fig. 3(b). Comparison of Fig. 3(a) and 3(b) shows clearly that the peak values that are attained by the residual disturbance in Fig. 3(b) exceed the peak values attained in Fig. 3(a). Accordingly, we conclude that the introduction of time delay in one of the signal paths, in order to enable the peaks of the impulse responses of the parallel paths to occur at the same time, may cause a deterioration in the impulse-noise-suppression performance of the scheme.

As a final illustration, consider the arrangement shown in Fig. 4, in which each block consists of an amplifier with a high-Q single-tuned load, the gain and bandwidth of each being those marked inside the block. In this example, the instantaneous amplitude of $N_o(t)$ is

$$E_N(t) = 2a A_N G_1 G_2 \left\{ (at) e^{-at} - \frac{bk}{k-1} \left[e^{-at} - e^{-kat} \right] \right\} \quad (27)$$

The function in braces is of the form

$$f(x) = xe^{-x} - \frac{bk}{k-1} f_1(x), \quad (28)$$

where $f_1(x)$ is the function defined in Eq. (11). The choice of b that results in equal peak values for the subtracted terms in Eq. (28) is

$$b = k^{1/(k-1)}/e \quad (29)$$

but the peak values are attained at different values of x . Note that $b \rightarrow 1$ as $k \rightarrow 1$, and that if $k = 1 + \epsilon$, $\epsilon \ll 1$, then $b \approx e(1 - \epsilon)$. The ratio of output signal amplitude to peak-noise amplitude is given by

$$\left[\frac{S_{\text{peak}}}{N_{\text{peak}}} \right]_{\text{out}} = \left[\frac{E_s}{2aA_N e^{-1}} \right] \frac{(1-b)/e}{[f(x)]_{\text{max}}} \quad (30)$$

The first factor on the right-hand side represents the ratio of signal amplitude to peak-noise amplitude at the output of the upper branch in Fig. 4. The improvement attributable to the subtraction is gauged by the factor

$$\left(\frac{1-b}{e} \right) / [f(x)]_{\text{max}}$$

Direct analysis (with the use of the method illustrated in the preceding example) shows that

$$\left(\frac{1-b}{e} \right) / [f(x)]_{\text{max}} < e/2; \quad \text{for } k > 1$$

and this upper bound is only 2.67 db. The same is true if $b = 1/k$.

VII. CONCLUDING REMARKS

The preceding examples show that it is possible to achieve substantial reductions in the relative strength of impulsive disturbances by means of simple linear-cancellation circuits. This technique can be combined with a preclipping operation of appropriately chosen threshold to effect almost complete elimination of the harmful effects of impulse noise, although some forms of the linear-cancellation technique will achieve such a result without the use of preclipping. The subtraction that results in these reductions is effective against a moderate number of impulses whose effects overlap completely or partially; it is not effective against very large numbers of randomly superimposed impulses (such as shot noise for example) because of the large number of residual overlapping disturbances caused by the input noise pulses. In this second situation, the coherent nature of the signal and the large number of the random residual noise disturbances combine to cause a decided deterioration in the mean signal to mean random-noise power ratio.

The use of delay in one branch to make peak values of envelopes coincide is not necessarily beneficial--it may degrade the performance achievable without the use of delay.

Some interesting ramifications of this technique, as well as specializations of it for specific modulation systems, such as cascading of simple linear-cancellation blocks, postdetection subtraction, and so forth, can be derived from the fundamental principles presented here, but this will not be attempted here. Experimental circuitry together with confirmational data, are also available.^{2,3}

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ACKNOWLEDGMENT

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(Office of Scientific Research, Air Research and Development Command), and the U. S. Navy (Office of Naval Research).

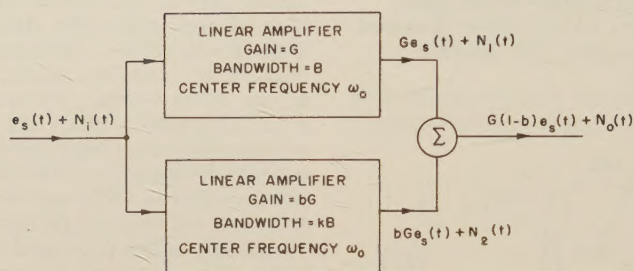


Fig. 1. Fundamental block diagram of the linear-cancellation technique for suppressing impulse noise.

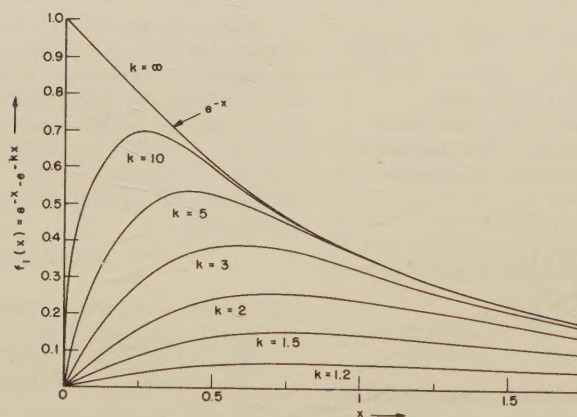


Fig. 2. Normalized impulse response of the arrangement of Fig. 1 when the paralleled amplifiers have single-tuned, high-Q loads.

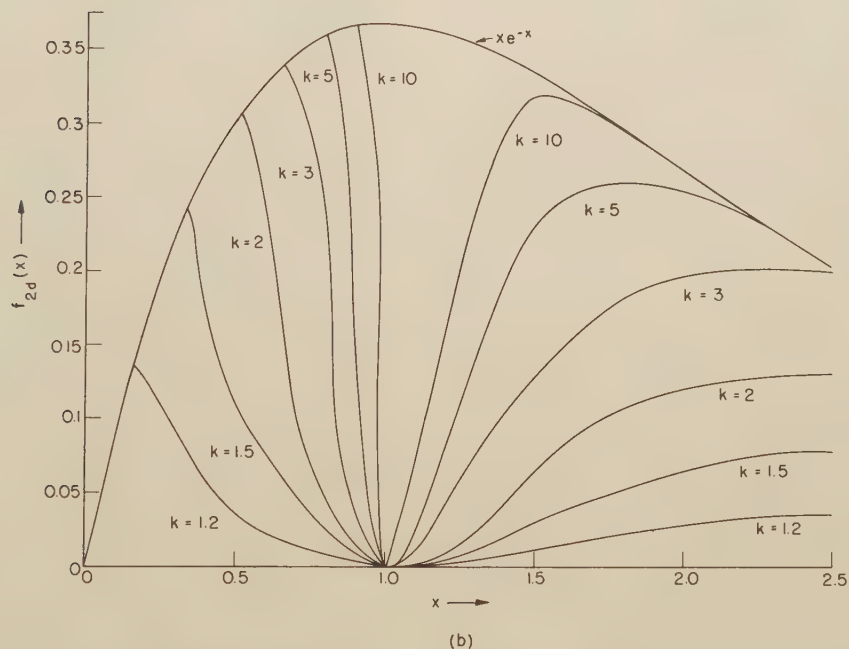
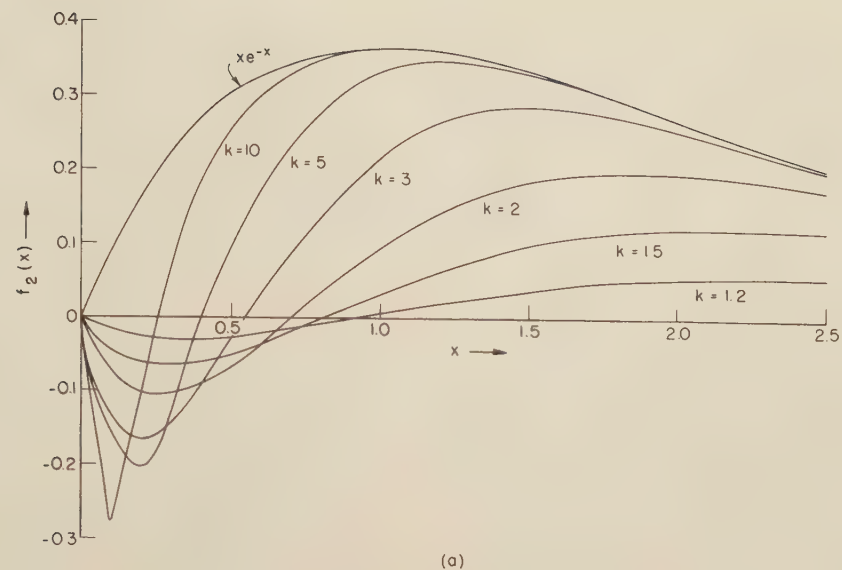


Fig. 3. Impulse noise envelopes described by : (a) $f_2(x)$ of (25);
(b) $f_{2d}(x)$ of (26).

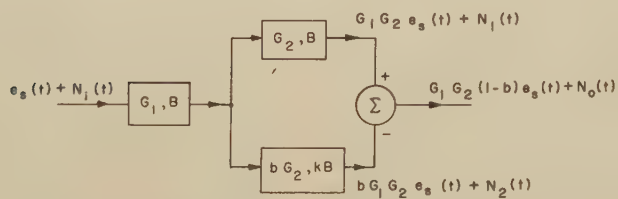


Fig. 4. Example used in illustrating the effect of narrow-band filtering the impulse noise before applying it to a simple linear subtraction scheme. (See (30) et seq., and compare with (14) et seq.)

THEORY AND PERFORMANCE OF VEHICULAR CENTER-FED WHIP ANTENNA

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Summary

The design considerations and results of extensive performance tests of a novel antenna for vehicular communication in the frequency range 30 to 76 mc are given together with applicable theory. The principal features of the antenna include feeding a whip antenna at its center through a coaxial cable, terminating it at its base into a variable reactor with different settings for each one of ten bands, providing separate built-in fixed-tuned matching networks for each band and switching them by means of automatic remote control which is activated by the frequency dial of the radio set. The design of the base reactor comprising a transmission line wound on a ferrite core having a tapped coil connected in parallel with it, and of the transmission line inside the whip are discussed in detail. Measured current distributions, radiation patterns, impedance plots and other performance data are presented.

Introduction

This article deals with a novel type antenna, referred to as the "center-fed whip," which was previously described, if only briefly and in general terms.^{1,2} Since then, the antenna has been developed and tested further, and the US Army has decided to replace the familiar base-fed whip antenna for vehicular tactical applications in the 30 to 76 mc frequency range with this new antenna. These facts by themselves might not be sufficient reason to elaborate here on its design and performance. It is believed, however, that the center-fed whip offers advantages in other applications, after suitable modification. A detailed account should be of more general interest also because it illuminates or brings to the fore poorly understood electrical phenomena related to ground based vertical antennas.

Principal Features

The external appearance of the center-fed whip shown in Fig. 1 differs little from the common base-fed whip. Referring to the schematic diagram of Fig. 2, its principal features can be summarized as follows:

(1) The matching problem is solved by dividing the frequency range in a number of fairly wide bands (up to ten), by providing a separate fixed-tuned network for each of the bands at the base of the antenna, and by switching to the proper matching network by means of automatic remote control activated by the frequency dial of the radio set.

Thus the feed cable from the antenna to the radio set can be of arbitrary length. No manual tuning is involved. The VSWR is kept smaller than 3:1 over the entire range.

(2) The vertical antenna is fed at or near its center through a coaxial cable and terminated at its lower end into a lumped reactor whose other terminal is connected to the vehicle body. Its reactance can be varied by shorting taps of a solenoid which is part of this reactor. A switch deck ganged with the stepping switch for the matching networks automatically selects the proper reactance value for each band. This permits controlling the current distribution along the lower section of the radiator below the feed point within wide limits. The criterion used in this particular case for adjusting this control was to place a current node at or near the lower end of the lower section in all bands. As a result, the distortion of the azimuth pattern and the loss caused by unintentional excitation of the vehicle body are minimized. Most important, the decoupling between antenna and vehicle body actually achieved is sufficient to make differences in impedance due to vehicle type, mounting location and ground conditions negligibly small. As a secondary effect, the power gain with respect to the field strength on the ground as compared to a base-fed antenna, is improved noticeably at the upper end of the frequency range.

(3) The proven mechanical advantages of the whip configuration are not compromised. There are no moving parts, telescoping sections, or lumped reactors in the whip itself.

The following account of the design considerations has been broken down into those of the lumped reactor at the base, the coaxial cable inside the whip, the matching networks and the whip itself. It is to be understood, however, that the design of these parts is intimately interrelated.

Design of Cable Choke

The essential component of the lumped reactor at the base is a section of coaxial cable wound on a ferrite core, called the "cable choke." It is this component which distinguishes the center-fed whip electrically from the well known "sleeve antenna." It is also the most critical component of the antenna. The outer conductor of the cable of the cable choke is connected at one end to the outer conductor of the cable inside the whip. At the other end it is connected to the metal container which protects and shields it, and which in turn is bonded to the vehicle body (see Fig. 2).

In such an arrangement, the cable choke constitutes an inductor with a fairly large capacitor in parallel formed by its self-capacitance and the capacity of the base insulator and other associated hardware. For the purpose of adjusting the reactance of this combination it is conceivable to tap the turns of the cable choke directly or to connect additional capacitors in parallel, as required for the adjustment. The first method is cumbersome for reasons of fabrication and introduces unnecessary losses. The second method means that additional energy is stored in the base reactor for a given voltage across it, and is, therefore, detrimental to the broadband properties of the antenna. It is for this same reason that it is important to keep the stray capacity of the base reactor assembly to a minimum. The method of adjustment actually employed is to connect a tapped cylindrical, air core coil with relatively loose coupling between turns in parallel to the cable choke and to short the taps of this coil, as desired. This method has the only disadvantage of requiring an inductance of the cable choke slightly higher than required for the other methods. Fig. 3 shows the tapped coil, called here the "shunt coil." The cable choke is located below it inside the neck of the housing and hidden from view.

For reasons explained later it is desirable to keep the length of cable used in winding the cable choke to a minimum. A ferrite core has been employed, therefore. Since the antenna is to handle a power of 70 watts, the strong increase of the losses in the ferrite with increasing induction had to be considered, in addition to the permeability of the ferrite body and the geometry of the core. The increase of the losses with frequency depends, of course, on the type of ferrite body, such as Q, Q₂ and Q₃ material. This increase with frequency turned out to represent no difficulty, for two reasons. One reason is that the voltage across the base reactor for constant antenna input power decreases with frequency, due to the increase of the radiation resistance. The measured base voltage is shown in Fig. 4. The other reason is that, for a given voltage, the RF current through the cable choke decreases with frequency, since the method of adjustment employed permits maintaining a constant inductance of the cable choke itself over the entire frequency range. As the frequency is increased the inductance of the shunt coil is lowered (in steps) and a larger portion of the circulating currents due to the capacity of the base assembly flows through it rather than through the cable choke. As a result, the losses in the base reactor assembly, using a Q₂-ferrite core, do not exceed about 5% of the antenna input power, as shown in Fig. 5. These losses have been determined by measuring the conductance G of the base assembly with the whip removed and the voltage V at the base under operational conditions, and computing V^2/G .

Thanks to the ferrite core, only four and a half turns of cable are necessary to obtain an inductance of the cable choke of $2\mu\text{h}$ or more,

depending on the core configuration. The length of the cable required is only about 23 inches. The ferrite core is a cylinder 1.0 inch in diameter and 2 inches long. The shunt coil is $3/4$ inches in diameter, $1\frac{1}{8}$ inches long and has 18 turns. The stray capacity in parallel to the cable choke and shunt coil is in the order of 20 pF.

Design of Coaxial Cable

The coaxial cable "inside" the whip and the cable choke is terminated into the impedance which the antenna represents at the center-feed point. This impedance can be measured only indirectly by measuring the input impedance of the cable at the "grounded" end of the cable choke and "transforming" this latter impedance to the feed point on the Smith Chart. It depends, of course, on the actual adjustment of the base reactor which will be discussed later. A representative set of frequency curves for the center-feed-point impedance in all 10 bands is given in the chart of Fig. 6. It can be seen from this chart that a characteristic impedance Z_0 of 125 ohms results in the lowest maximum SWR in the cable over the frequency range obtainable. Because of the large spread of the impedance values, a different Z_0 would merely lower the SWR at one end of the frequency range and raise it at the other end. A high SWR in the order of 17:1 is thus unavoidable, at least if the possibility of placing matching devices at the feed point is ruled out for mechanical reasons. It is possible, however, to reduce the SWR somewhat in a portion of the cable by making a transition to a lower Z_0 at a certain distance below the feed point. This possibility is illustrated in the chart of Fig. 7 which shows the impedance obtained by transforming the feed-point impedance through a 39 inch section of 125-ohm cable and then normalizing it to 73 ohms. Thus, if the Z_0 of the adjoining section of feed cable is changed to 73 ohms, the SWR in the adjoining section is only 11:1. Such a transition has the added advantage that it facilitates the problem of broadband matching of the resulting impedance at the input terminal to 50 ohms. It has not yet been incorporated, however, into the models used in performance tests.

Nevertheless, the SWR is so high that a close check on the cable losses is warranted in view of the reduction in antenna efficiency caused by them. It would seem that such a check could easily be made by calculating the losses according to standard practice but it turns out that the formulae for transmission line losses found in textbooks³ are not accurate enough in this case. The reason is that these formulae assume line lengths of many wavelengths and neglect certain terms which are usually insignificant under this assumption. In the case of the center-fed whip, however, the line length is in the order of a quarter to a half wavelength only. A derivation without the usual neglect yields the following ratio of the line loss P_1 to the power P_2 delivered to the terminating impedance $Z_e = R + jX$:

$$\frac{P_L}{P_e} = e^{2\alpha l} \left[1 + \frac{|T|^2}{1-|T|^2} (1 - e^{-4\alpha l}) \right] - 1 - 2\alpha l A \quad (1)$$

In this formula, T is the reflection coefficient of the terminating impedance,

$$\Gamma \equiv |T| \cdot e^{j2\psi} = \frac{Z_e - Z_0}{Z_e + Z_0} \quad (2)$$

α is the attenuation constant, β the phase constant of the line and A an abbreviation for

$$A = 2 \frac{\sin \beta l}{\beta l} \left[\frac{\cos \beta l |T| \cos 2\psi + \sin \beta l |T| \sin 2\psi}{1 - |T|^2} \right] \quad (3)$$

which can also be written:

$$A = \frac{|T|}{\beta l} \left[\frac{\sin(2\beta l - 2\psi) + \sin 2\psi}{1 - |T|^2} \right] \quad (4)$$

When expressed in terms of normalized R and X , this reads:

$$A = \frac{\sin \beta l}{\beta l} \left[\cos \beta l \frac{\left(\frac{R}{Z_0}\right)^2 + \left(\frac{X}{Z_0}\right)^2 - 1}{2\left(\frac{R}{Z_0}\right)} + \sin \beta l \left(\frac{X}{R}\right) \right] \quad (5)$$

In the case $2\alpha l < 0.2$ neper (or 1.74 db), it is in a good approximation:

$$\frac{P_L}{P_e} \approx 2\alpha l \left[\frac{1 + |T|^2}{1 - |T|^2} - A \right] \quad (6)$$

or, expressed in terms of R and X :

$$\frac{P_L}{P_e} = 2\alpha l \left[\frac{1 + \left(\frac{R}{Z_0}\right)^2 + \left(\frac{X}{Z_0}\right)^2}{2\left(\frac{R}{Z_0}\right)} - A \right] \quad (7)$$

If the line is an integral multiple of a half wave long, $A=0$. For lines which are so short that $\sin \beta l \approx \beta l$ and $\cos \beta l \approx 1$, it is:

$$\frac{P_L}{P_e} \approx 2\alpha l \left[\frac{Z_0}{R} - \beta l \left(\frac{X}{R}\right) \right] \quad (8)$$

It is interesting to note that the losses for an inductive load are always smaller than for a capacitive load in this case!

When these formulae were applied to the case of the center-fed whip it turned out, surprisingly, that the cable losses of any of the types of cables considered never amount to more than 0.2 db, in spite of the high SWR, and are, therefore, negligible. Obviously, the short line length and the low attenuation constant are the dominating factors.

The high SWR is more of a design problem with respect to the power handling capability of the cable. Again, the situation here is different from the usual case of a transmission line which is electrically more than a half wavelength long and, therefore, carries at least one voltage maximum and one current maximum. In the case of the center-fed whip the cable is only about a quarter wavelength long at the low end of the frequency range. Either a voltage maximum or a current maximum will occur then, depending on the phase angle of the reflection coefficient. A voltage maximum is practically of no concern since the voltage handling capability of most types of cables corresponds to a higher power than their current handling capability. If a current maximum actually occurs in the cable, its power handling capability under this condition, as compared to that for a matched load, is not simply computed by multiplying it with a factor equal to the inverse of the SWR, but by multiplying it with a somewhat larger factor. This is due to the non-uniformity of the current distribution along the cable and the equalizing effect of heat conduction etc. The situation is so complex that a near-optimum design is difficult to obtain. Yet, it is important to approach it because the broadband properties of the cable choke and the mechanical properties of the whip are adversely affected by a large cable diameter.

In connection with these difficulties, the Crosley Division of the AVCO Corporation in Cincinnati, Ohio, made an interesting proposal while developing the center-fed whip under contract with the US Army Signal Research and Development Laboratory. The proposal has been adopted in the final design. AVCO proposed to replace the coaxial cable of the "cable choke" with a bifilar winding. This modification permits increasing the current handling capability of the winding as a transmission line feeding the antenna, without increasing the self-capacitance and size of the cable choke. It introduces, however, a certain amount of leakage of rf energy from this transmission line which couples into the antenna at its base. No detrimental effects attributable to this leakage have as yet been observed, though.

Matching Networks

All components of the antenna have been designed with a view towards mass production. The matching networks are, therefore, of simple circuitry. The circuits of some of them are shown in

Fig. 2. The SWR in the cable from the radio set to the antenna referred to 50 ohm does not exceed 3:1 in any of the 10 bands. It would have been possible to reduce the number of bands to six or less, but only at the expense of more complex and costly networks. The photograph of Fig. 3 shows the arrangement of the networks inside their housing, together with the 3-deck 10-position step switch is remotely and automatically controlled by the setting of the frequency dial of the radio set by means of a solenoid actuator. The losses in the matching networks are, for practical purposes, negligible.

Whip Dimensions

The designer of a base-fed whip for a given broad frequency range has a choice with respect to only two important parameters which bear on the electrical performance: The total length and the thickness. The center-fed whip configuration offers two additional degrees of freedom in designing: The reactance of the base reactor assembly for each band, and the ratio of the length of the lower section to the total length, or, in other words, the height of the feed point.

This greater freedom makes it possible to attain better performance or to pursue special objectives, or both. But it also complicates the task of arriving at optimum dimensions. Finding them empirically, that is by means of the "cut-and-try" method is much more tedious than in the case of the base-fed whip. In the initial phase of the development an attempt was made, therefore, to arrive at near-optimum dimensions on the basis of theoretical considerations. A number of simplifying assumptions were made to this end, such as sinusoidal current distribution in each of the two sections of the whip, perfectly conducting plane ground and placement of the lower end of the antenna flush with the ground plane. Two different adjustments of the base reactor were investigated: One which yields maximum directivity gain for the horizontal direction, and one which yields a node of the current distribution at the base. The latter condition was investigated because it was expected to result in low coupling between antenna and vehicle body. The pertinent formulae for the vertical radiation pattern, radiation resistance etc. are available in the literature⁴ to which reference is made. They were evaluated numerically with the help of a computer for several sets of dimensions, each time for a number of frequencies distributed over the range.

In the light of the experimental results obtained later these calculations proved useful only as a guide. The assumptions mentioned are not sufficiently realistic to permit making accurate predictions. More realistic assumptions would make the calculations hopelessly cumbersome.

The total length adopted in the latest stage of development is 122 inches. A shorter length is electrically inferior or difficult as explained in the following paragraph, and a greater length is deemed undesirable for mechanical reasons in tact-

ical military applications. It is likely however that the optimum length from the sole viewpoint of electrical performance is greater than the adopted length. It is interesting to note that a length of 10 feet is more than $3/4$ of a wavelength at the upper end of the frequency range and thus well above the optimum length of a base-fed whip which is somewhat less than $5/8$ of a wavelength, depending on the ground system etc.

The effect of varying the height of the feed point while maintaining the total length has not yet been investigated. It is, therefore, not known how close the adopted height of 53 inches comes to the optimum height.

Recent advances in molding plastic whips have permitted making the diameter relatively large without sacrificing mechanical performance. It should be mentioned in this connection that the electrical difficulties caused by decreasing the diameter are more severe in the case of the center-fed whip than in that of the base-fed whip.

A great deal of thought has been given to the design of the base support consisting mainly of a plastic insulator and a steel spring. It is important from the electrical viewpoint in as much as it affects the stray capacity in parallel to the cable choke. Adaptability to installation in different type vehicles is another important consideration. The design has not yet been finalized, however.

Actual Adjustment and Performance

The criterion used in adjusting the taps of the coil connected in parallel to the cable choke was primarily the position of the node of the current distribution along the whip. Specifically, this node was placed at or near the base insulator, unless other considerations suggested a different adjustment. This principle of adjustment was based on the tentative assumption that the coupling between whip and vehicle body is a minimum for this position of the current node. The measured current distribution after the adjustment had been made is shown in Fig. 8. The measured susceptance of the base reactor is shown in Fig. 9. Evidently, a node at the base insulator corresponds to zero susceptance.

This principle was compromised, however, both at the lower and upper end of the frequency range. It was found extremely difficult to obtain a match to 50 ohms at the input of the matching networks at the lower end of the frequency range when the current node was positioned at the base insulator. This finding is explained by the fact that the 10-foot whip is only 0.3 wavelength long at 30 mc and has, therefore, a low radiation resistance and a high reactance under these conditions. By positioning the (virtual) current node below the base, corresponding to a capacitive susceptance of the base reactor, the difficulty of matching was greatly alleviated (see Fig. 8, Band #1 and 2). This compromise proved to be acceptable from the point of view of decoupling, judging by the uniformity

of the horizontal radiation pattern shown in Fig. 10.

At the upper end of the frequency range, it was found empirically that the maximum power gain with respect to the field strength on the ground occurred for an adjustment of the shunt coil which made the base reactor inductive. In this frequency band, the current distribution along the whip in itself and its dependence on the base reactance cannot be explained in terms of simple transmission line theory (see Fig. 8, Band #10). A more systematic investigation is necessary to find a theoretical explanation. The adjustment of the shunt coil for maximum power gain did not lead to a noticeable excitation of the vehicle body, as evidenced by the uniformity of the horizontal pattern shown in Fig. 11. This observation agrees with the fact that the measured current distribution shows a relatively small current at the base insulator (see Fig. 8).

A particularly striking and unexpected feature of the measured current distribution is the rapid rise of its center of gravity near the upper end of the frequency range. It is believed that this rise is at least part of the explanation for the observed increase in power gain of 6.2 db in Band #10 (71 to 76 mc) and of 2.2 db in Band #9 (65 to 71 mc), as compared to a base-fed whip (of any length). The difference in power gain between the two antennas in the other bands (30 to 65 mc) did not exceed ± 1 db and is considered insignificant in view of measuring inaccuracy.

The power gain comparisons referred to above have been made at relatively close range (about 150'). The results of large scale field tests have recently led the author to suspect that the power gain comparison of antennas with different vertical radiation patterns at close range differs from one made at large distances, say, 10 or 15 miles away. Such a phenomenon can occur if refraction of the electromagnetic waves in the atmosphere near the ground affects the field strength on the ground, and if the vertical radiation patterns of the antennas under comparison differ at low elevation angles. Strong evidence for the existence of such effects in the standard broadcast frequency range (.6 to 1.5 mc) has recently been obtained. A search for a similar effect in the 30 to 76 mc range is planned.

Fig. 12 is a sample of the input impedance versus frequency measured at the input of the matching network for one of the bands. It illustrates that the variation due to the type of vehicle used is small.

Conclusions

The considerations involved in the design of a vehicular center-fed whip antenna for a certain tactical military application have been described in sufficient detail to permit their use in other applications. The performance measurements given show the superiority of the center-fed whip over the common base-fed whip, in particular the reduced coupling between antenna and vehicle body. Differences due to vehicle type, mounting location and type of surface below the vehicle are negligibly small which is of paramount importance in this application. Certain aspects remain which have not yet been investigated or are poorly understood.

Acknowledgement

Contributions to the successful development of the center-fed whip have been made by numerous personnel of the US Army Signal Research and Development Laboratory, Fort Monmouth, N. J. and the Crosley Division of the AVCO Corporation, Cincinnati, Ohio. The experimental investigations mentioned have been conducted mainly by W. P. Czerwinski and D. Campbell of USASRD.

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Fig. 1. Typical installation of center-fed whip antenna on 3/4-ton truck.

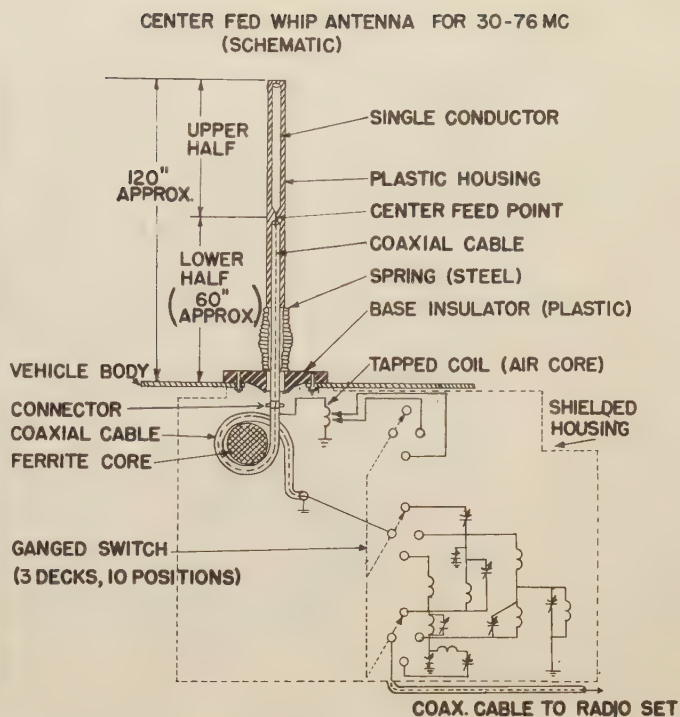


Fig. 2. Schematic diagram of center-fed whip.

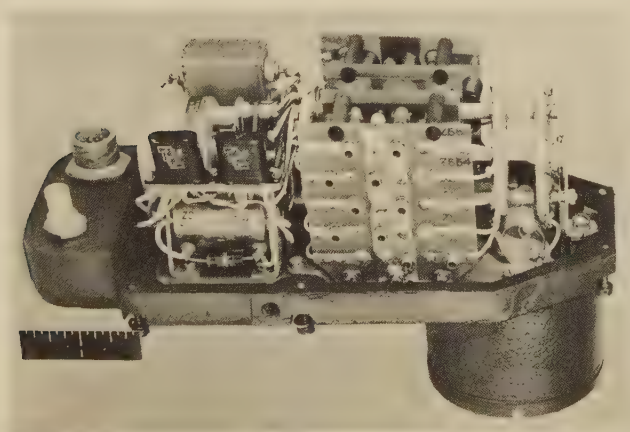


Fig. 3. Matching unit of center-fed whip with lid removed. Bottom view showing shunt-coil (left), matching networks (center), solenoid actuator (right) and 3-deck, 10-position step-switch.

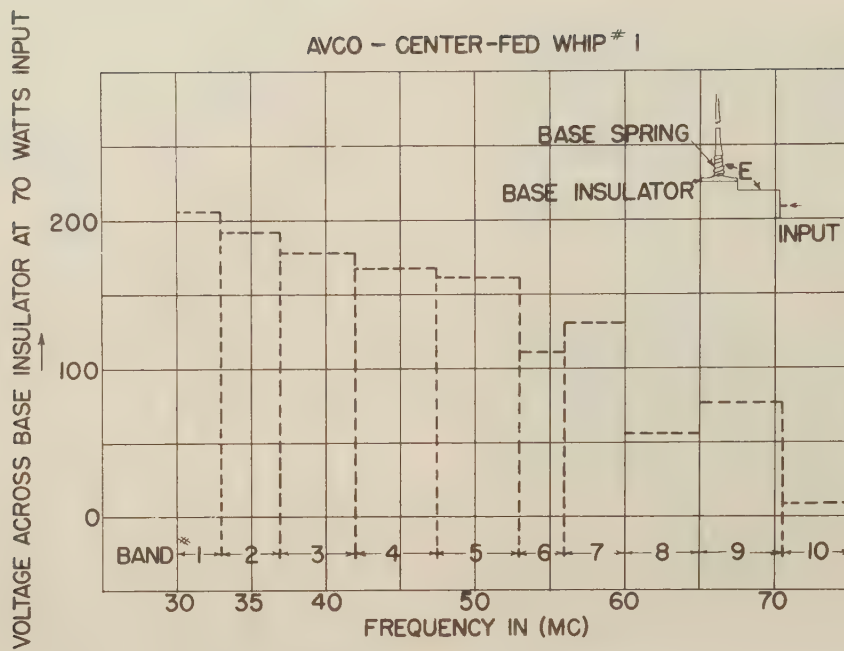


Fig. 4. Voltage across base-reactor vs frequency for 70-watt input power.

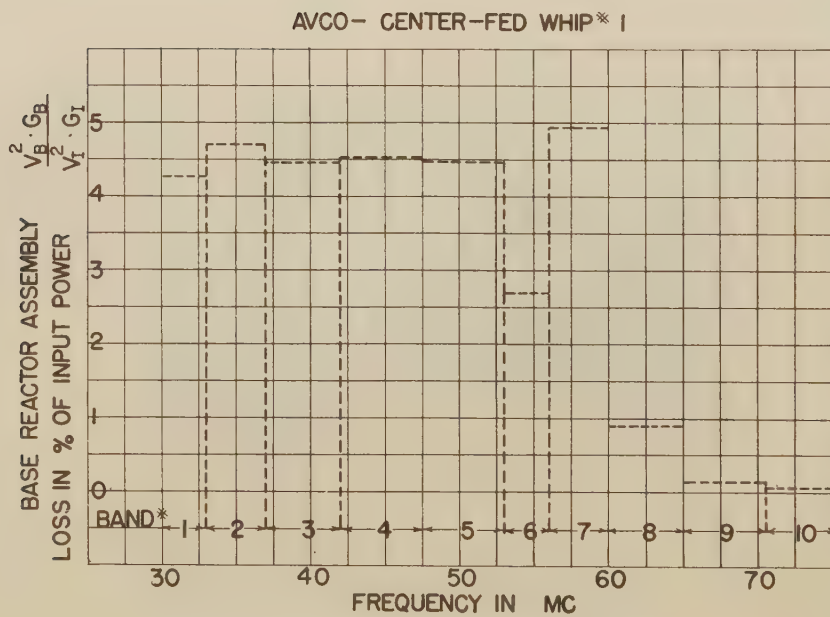


Fig. 5. Losses in base-reactor assembly vs frequency.

AVCO CENTER-FED WHIP FEED
POINT IMPEDANCE (SER.# 20)

NORMALIZED TO 125 Ω

VSWR MAX = 17 TO 1
AT 60 MC/S

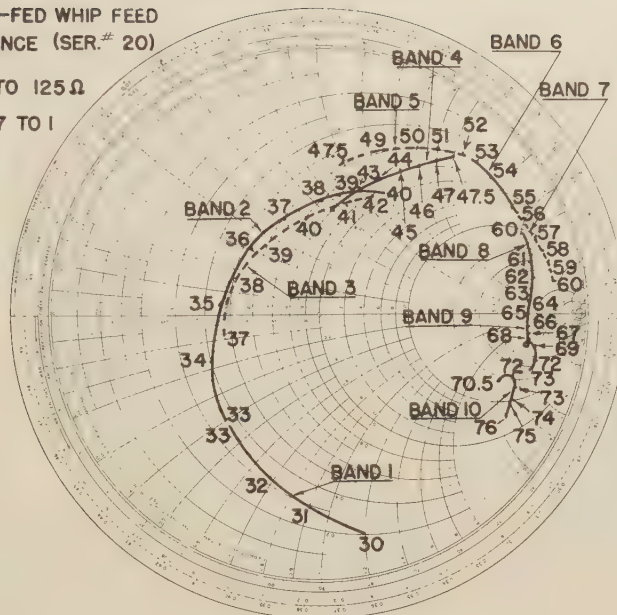


Fig. 6. Antenna impedance at feed point of center-fed whip vs frequency, referred to 125 ohms.

AVCO CENTER-FED WHIP (SER.# 20)
FEEDPOINT IMPEDANCE
TRANSFORMATION

VSWR MAX = 11 TO 1
AT 60 MC/S
(BAND 7)

TRANSFORMED
THROUGH
0.995 METERS
OF 125 Ω CABLE ($\frac{V}{C} = .84$)
AND THEN NORMALIZED TO 73 Ω

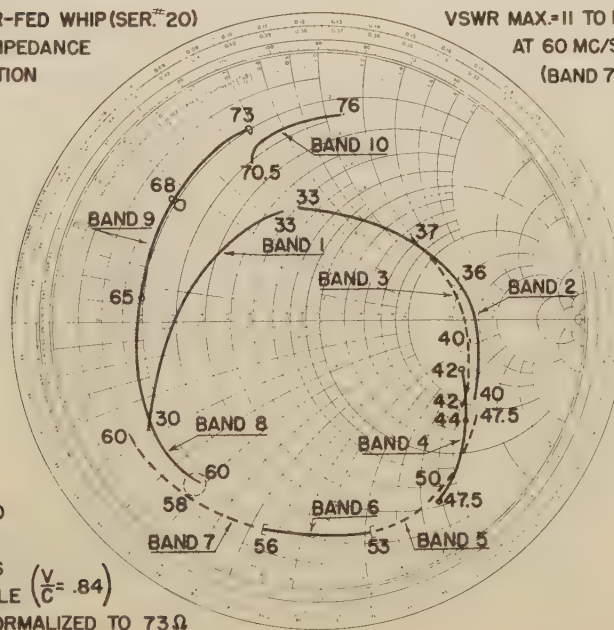


Fig. 7. Antenna impedance of Fig. 6 after transformation through a 39-inch long section of 125-ohm coaxial cable, referred to 73 ohms.

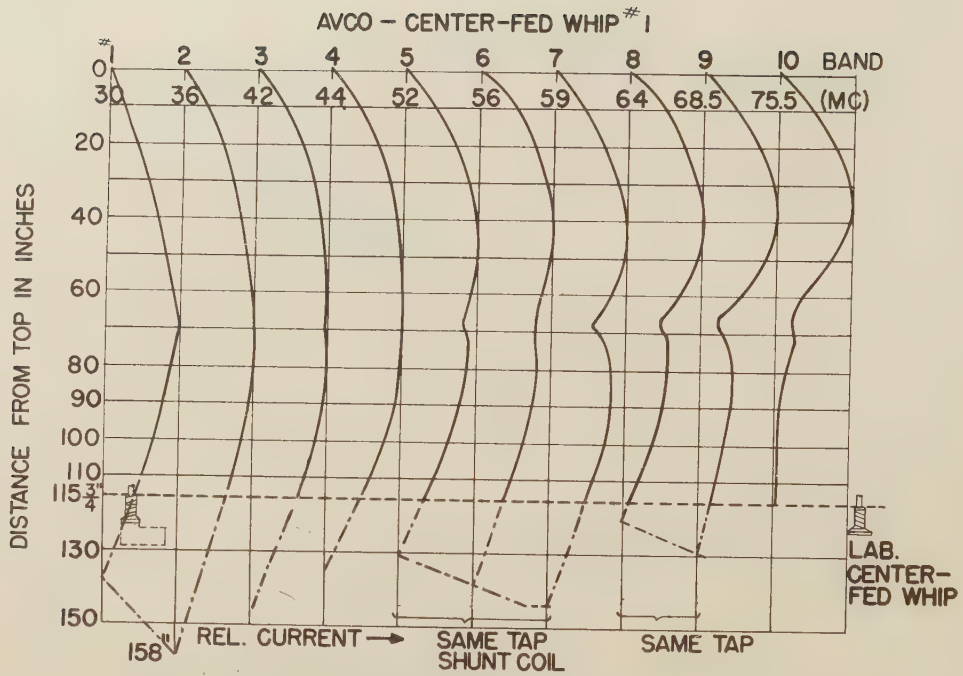


Fig. 8. Current distribution along center-fed whip for several frequencies, one for each band.

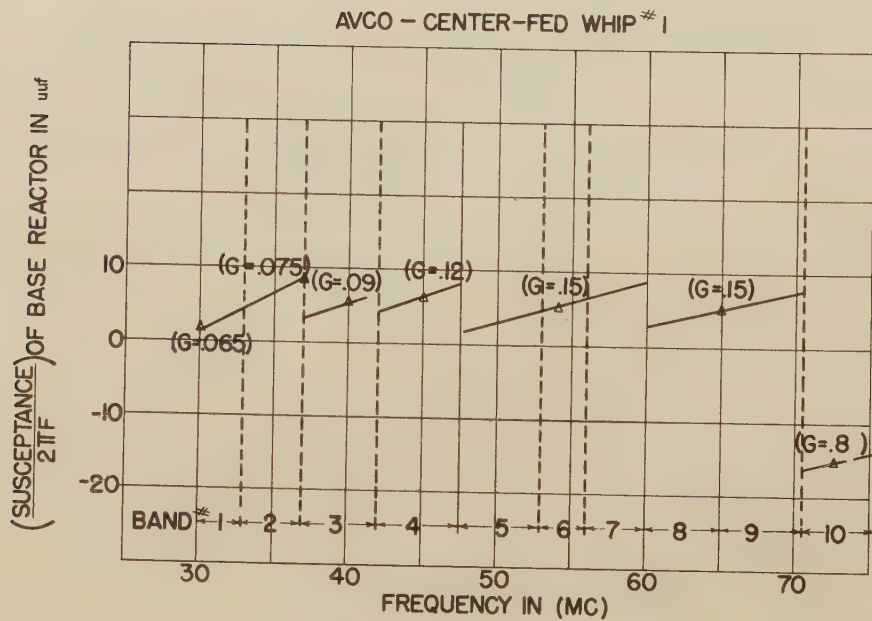


Fig. 9. Measured susceptance of base reactor vs frequency.

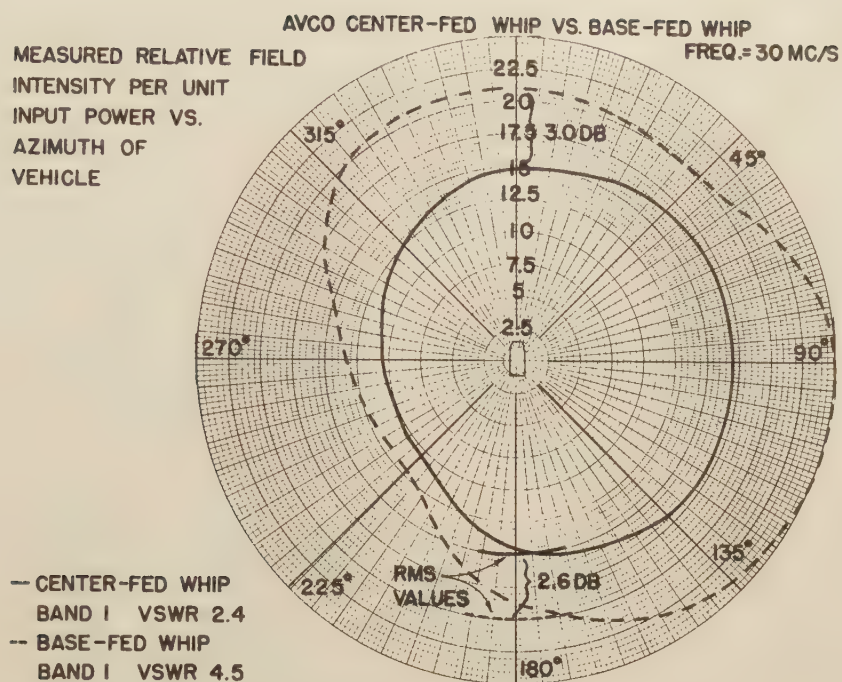


Fig. 10. Horizontal radiation pattern at 30 mc for center-fed and base-fed whip.

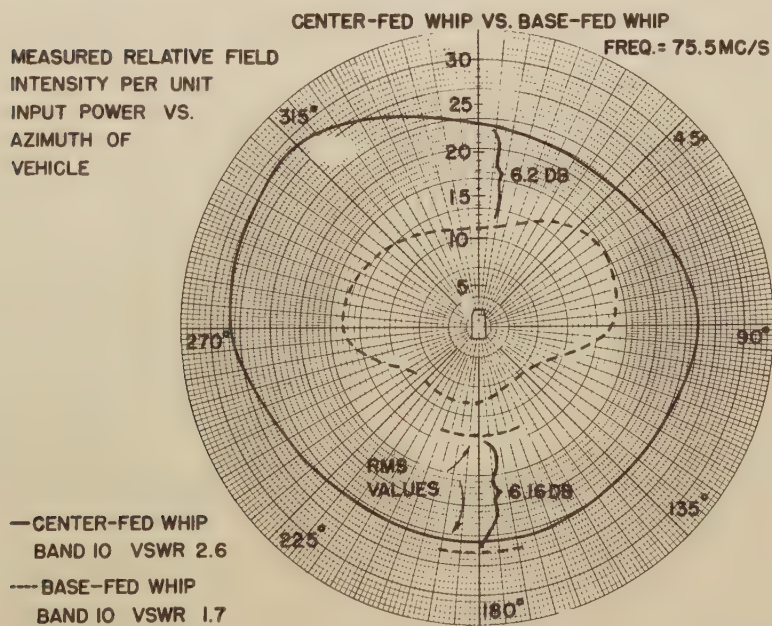


Fig. 11. Horizontal radiation pattern at 71 mc for center-fed and base-fed whip.

SAMPLE OF INPUT IMPEDANCE
WITH MATCHING NETWORK
FOR VARIOUS TYPE
VEHICLES

EXPERIMENTAL
MODEL

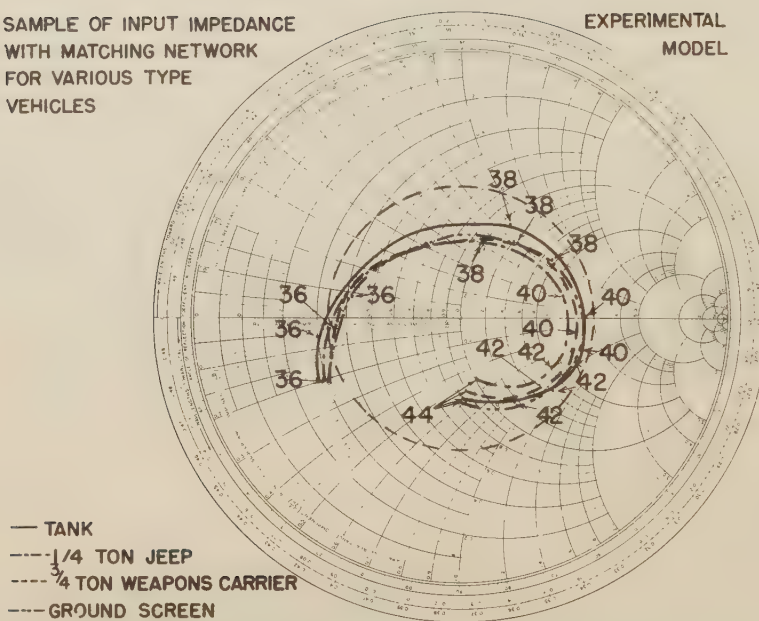


Fig. 12. Sample of input impedance vs frequency plot. Measured at input of matching unit for band #4, when center-fed whip mounted on 3 different type vehicles.

A BROADBAND 160-MC COLINEAR ARRAY

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ABSTRACT

The design of a 3-element colinear array is described. Over the 152-162 MC band, the array has a minimum gain of 4 db over a half-wavelength dipole and VSWR less than 1.5. The circularity of its horizontal pattern is $\pm 1/4$ db. The array elements are covered and supported by a Fiberglas tube. DC grounds are provided for these radiating elements for lightning protection and reduction in static noise.

INTRODUCTION

The array consists of three colinear half-wavelength dipoles which are covered and supported by a Fiberglas tube as shown in Fig. 1. The dipole elements are made of a series of metallic cylinders except for the lower half of lowest dipole where four slanted ground rods are used. The connecting harness is arranged coaxially with the dipole-elements. The isolation between harness and dipole-elements is achieved by using quarter-wavelength chokes built into these elements. In this manner, the effect of harness on radiation characteristics is eliminated to a great extent.

The R. F. resonance of a dipole occurs when the length of dipole is near to but less than half wavelength of the applied R. F. signal. The sharpness of the resonant curve, i. e. current amplitude vs frequency, reduces with increasing dipole diameter. Thus broadband dipole usually has a relatively large diameter which is generally expressed in terms of wavelength or dipole length (1). In practice, the dipole diameter is dictated by limitations of weight and windage, especially in the case of a vertical antenna array. Once these mechanical design goals are set, maximum bandwidth may be obtained by placing the dipole radiating surfaces outmost within the given array diameter. This criterion has led to the design as shown in Fig. 1.

IMPEDANCE BANDWIDTH

The lines connecting top and bottom dipoles are made of low-loss foam-dielectric cable with a same size of RG-11/U. The characteristic impedance of this cable is 75 ohms which presents a better match to dipole elements. Thus harness frequency-sensitivity is re-

duced. These lines are each one wavelength long electrically and thus three radiating elements are in fact fed with equal in-phase voltages. The reactance of this array is partially cancelled by the half-wavelength open-circuited shunt stub and the final match is achieved by a series quarter-wavelength transformer as shown in Fig. 1. The VSWR curve of this array is shown in Fig. 2 and it is seen that VSWR on a 50- Ω line is under 1.5 over the 152-162 MC band.

GAIN AND RADIATION PATTERN

The self-impedances of dipole-elements are first adjusted nearly to a same value. The mutual impedance of the middle dipole element is different from that of the end ones. However, the difference is small and may be neglected in the initial adjustment. Since the applied voltages are equal and in-phase, the currents in the dipole-elements of same input impedance are also equal and in-phase. Thus maximum gain is obtained.

R. F. coupling between a vertical antenna and its supporting metallic mast causes R. F. current flow in the mast which in turn causes gain reduction as a result of radiation in unwanted directions. To minimize this troublesome coupling, a combination of slanted ground rods and a quarter-wavelength choke immediately below as shown in Fig. 1 is used. The ground rods reduce quite effectively the mutual coupling and also help to reduce the overall array length. The choke below these four ground rods reduces local coupling and also serves as the mounting stem for the array. For best pattern characteristics, the slant angle of ground rod was found to be about 30° below horizon.

R. F. couplings between the outer conductor of harness and dipole elements are often troublesome. If the induced currents in the harness outer-conductor are in phase with the currents in the dipole-elements, all is fine. But there is little increase in gain as there is no increase in array aperture. However, if they are out of phase, the gain of array reduces drastically. Since the phase of the induced currents depend on that of the induced voltage and the impedance of harness outer conductor, the in-phase

condition rarely exists. Even if it does, it would be very frequency sensitive. Therefore, it is always wise to avoid such R. F. couplings. In this array such couplings due to local currents are reduced substantially by using quarter-wavelength chokes. The induced voltage is small as the harness is placed coaxially with, and is quite a bit thinner than, the dipole elements. The wave velocity in the harness is about 81% of light velocity and the physical length of one-wavelength-long cable is 0.81λ . The length of dipole is roughly 0.46λ . Thus the length of the harness outer-conductor between tips of adjacent dipoles is 0.35λ long. Since this length is far below R. F. resonance and the induced voltage in the harness section through mutual induction is small, the induced current flow on harness is small. It is in fact too small to effect the array radiation pattern.

Theoretical gain of a colinear array of 3 half-wavelength dipoles spaced 0.81λ center-to-center is about 4.5 db over a half-wavelength dipole (2). The gain of this three-element array measured by pattern integration is 4.1 db over a half-wavelength dipole over the band 152-162 MC as shown in Fig. 3. The maximum gain of a three-element colinear array is about 5.3 db when dipole is spaced about one wavelength center-to-center (2). In this case, the array is to be about 20% longer and care of preventing harness resonance must be exercised.

Typical vertical-plane radiation pattern of this array is shown in Fig. 4.

The beamwidth between the two half-power point is about 22° . The pattern circularity in the horizontal plane is within $\pm 1/4$ db.

MECHANICAL ASPECT

The outer conductor of harness is replaced by heavy wall tubing in actual design to provide low DC ground resistance which is essential for lightning protection and reduction in static noise. The array is 2-1/2 inches in diameter and 13-1/2 feet in length. The top element is detachable and the array length in shipping may be made slightly less than 12 feet. It weighs about 32 lbs. It is designed to stand a windage of 30 PSF with 1/2 inch radial ice.

ACKNOWLEDGEMENTS

The authors are grateful to Mr. J. S. Brown, Director of Engineering, ANDREW Corporation, for his guidance and encouragement in this work.

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- (2) R.F.H. Yang, "Optimum Element Spacing of Uniform Broadside Arrays", Wireless Engineering, April, 1955.

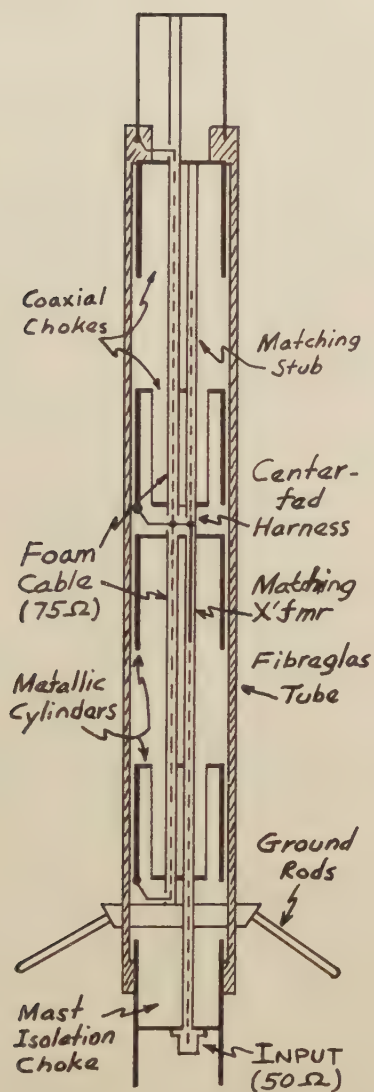


Fig. 1. A broadband colinear array.

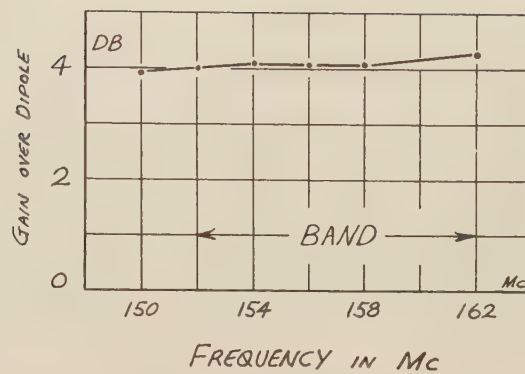


Fig. 3. Gain vs frequency.

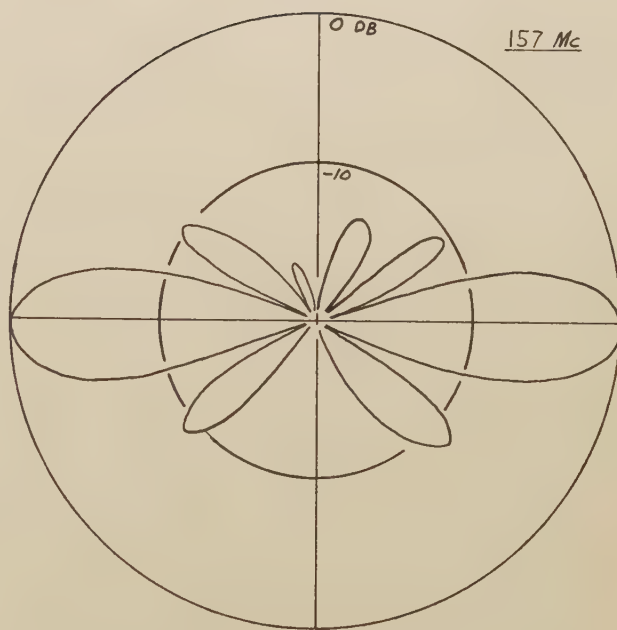


Fig. 4. Typical vertical pattern.

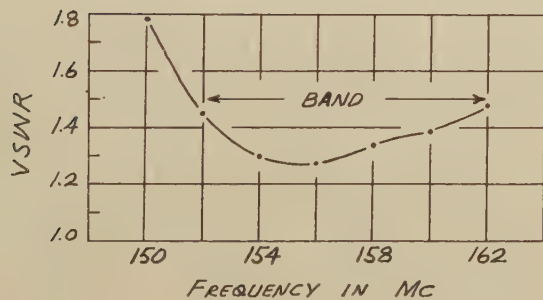


Fig. 2. VSWR vs frequency.

EFFECTS OF TOWER AND GUYS ON PERFORMANCE OF SIDE-MOUNTED VERTICAL ANTENNAS

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ABSTRACT

The omni-directional pattern of a vertical antenna when side-mounted on a mast or tower, is distorted as a function of tower diameter and separation in wavelength. The pattern is further effected by obstruction of metallic tower guys. Laboratory and field measurements of these effects are analyzed. Possibility of taking advantage of these effects for special coverage is suggested. Tower effect on antenna impedance is briefly discussed.

INTRODUCTION

Because of economy and certain operational necessity, many communication antennas are installed on a common supporting mast or tower. Since there is only one tower top, many of these antennas have to be mounted on the sides of the tower. Thus tower effect on coverage or horizontal pattern of these vertically polarized antennas becomes an important factor in the installation of a modern base station antenna system. The VSWR of a vertical antenna is also effected by the presence of tower. In the case of a guyed tower, the guys also effect the horizontal pattern. This paper reports the results of some experimental studies on these effects. Experimental work was done by scale model technique as well as field measurement of actual antenna system.

The radiation pattern of a short dipole near a infinitely-long conducting circular cylinder has been treated thoroughly by Carter (1). In our problem, however, tower is not a circular cylinder nor is it infinitely-long. Nevertheless Carter's work is still very useful in that it gives a first order of approximation and indications as to what should be studied in an experimental program. In general, it may be said that resultant pattern depends on cylinder diameter and distance between dipole and cylinder measured center-to-center. The radiation intensity to the rear of dipole reduces with increasing cylinder diameter. As the distance between dipole and cylinder increases, the number of maximum and minimum in the horizontal radiation pattern increases but with improved circularity.

In the experiments performed on the tower effect on horizontal radiation pattern, both dipole and folded-dipole were used as a matter of convenience.

Experimentally no difference between the two has been found as long as the distance in the latter case is specified from tower center to the center between the fed and the parasitic elements of a folded dipole. Theoretically it may be proved that the effect of long conducting cylinder on the radiation pattern of a vertical colinear array is same as that of a dipole. This has also been verified experimentally. In the following, we shall refer these antennas as vertical antennas.

TOWER EFFECT ON HORIZONTAL PATTERN

If the spacing between a vertical antenna and its supporting tower or mast is less than $\lambda/4$, the pattern is cardioidal with maximum intensity in the direction in which the antenna is mounted away from the mast. If mast diameter is less than $\lambda/8$, the front-to-back ratio varies from 3 to 10 db as the mast diameter varies from $.01\lambda$ to $\lambda/8$ as shown in Fig. 1. As it can be seen from the curve, the separation has small effect as long as mast diameter is small.

The pattern becomes bi-directional as the spacing is increased to about $\lambda/2$, and tri-directional for $3/4 \lambda$ spacing. Typical patterns are shown in Fig. 2. In these two cases where mast diameter is about $\lambda/16$, the pattern circularity is about ± 3 or ± 4 db.

As mast diameter becomes large in terms of wavelength, the general trend of spacing effect on pattern remains same as mentioned above. However, the pattern nulls become deeper with larger mast diameter. Measured patterns as shown in Fig. 3 indicate that the nulls to the rear are about 18 to 20 db down for $0.46\lambda \times 0.46\lambda$ square tower and 14 to 16 db for $0.23\lambda \times 0.23$ square tower, and that pattern is cardioidal for 0.33λ spacing and bi-directional for 0.52λ spacing.

For mast or tower with diameter in excess of $\lambda/4$ or so, omni-directional pattern may be obtained by two or more antennas of same kind fed in phase and mounted on different corners of a tower. If there is no space limitation, each of these antenna elements may be further mounted one wavelength apart vertically to form an vertical array. Fig. 4 shows the case of a triangular tower with each side measured 0.085λ wide. The pattern circu-

larity is improved from ± 5 db to ± 3 db with two vertical antennas and to ± 1 db with three antennas.

Fig. 5 shows the same effect for the case of a 0.38λ triangular tower. In this case the improvement is from ± 7 db to ± 3.5 db with three antennas. The poorer circularity is due to larger tower as compared to the previous case shown in Fig. 4.

Fig. 6 shows that as the tower sides become wider than a wavelength, there is no simple way to improve the pattern circularity except more dipoles are added so that the spacing between adjacent dipoles is made less than half of a wavelength.

There are two interesting points which are worthy to mention. (1) For a given tower, the pattern circularity appears to be better if the vertical antenna is mounted on broadside of a tower than it is mounted on the tower corner as it may be seen from the two patterns to the left of Fig. 5. (2) The pattern circularity of three vertical antennas equally spaced on a circle is somewhat independent to their relative orientation to a triangular tower as it may be seen from the two patterns to the right of Fig. 5. Since circularity improves as the diameter of this circle or the spacing between adjacent dipoles decreases, broadside mounting of dipoles is likely to be preferred.

EFFECT OF GUYS ON HORIZONTAL PATTERN

The effect of metallic tower guys on horizontal pattern is shown in Fig. 7. In the scale model a set of three guy wires was tied to the top of tower as shown. Radiation intensity was found to be about 3 db down in the direction of top dipole and 2 db down in the direction of bottom dipole. The difference was due to the different distances between the guy wires and the dipoles. The guy wire closest to the top dipole had greatest effect on horizontal pattern in that direction.

The guy wire effect on radiation pattern may be eliminated by inserting at the tower end either non-conducting robe about one or two wavelength long or break-away insulators. In the latter case, it was found from scale-model measurement that the guy wire effect was practically nil when all wires were broken electrically into three quarter-wavelength sections with four insulators near to and at the tower end. This indicates that the rest of the guy wires although very much conducting, are sufficiently far away from

the vertical antennas so that no significant RF current is excited to distort the radiation pattern.

FIELD MEASUREMENT

Fig. 8 shows the results of field measurement made on an array of four side-mounted folded-dipoles spaced one wavelength apart in elevation with an arrangement as shown in the inset of Fig. 7. The operating frequency was 43 MC. The tower was about 300 feet high, triangular in cross-section with face measured about 30 inches wide and guyed, beginning at tower top, by several sets of three steel robes without break-away insulators. In Fig. 8 the thin circles super-imposed on the map are the relative field intensity scales in db. The curve in heavy line is the measured pattern of scale model of this antenna system and it is the same curve as shown on the left of Fig. 7. The dotted circle is the pattern of a coaxial dipole mounted on the tower top with its effective height adjusted to that of the four-element side-mounted array. The gain of this array was measured by comparing the signal intensities received at three locations ranging from 40 to 73 miles away from station as marked by the black dots on the map while station transmitter was switching back-and-forth from the dipole to the array. In this manner, the propagation effect on measurement accuracy was eliminated. The measured values of gain are plotted as X in Fig. 8. Comparing to laboratory scale-model pattern, the gain was 1 db too low in the directions marked 73 miles and 52 miles and 2 db too high in the direction marked 40 miles. It is believed that this agreement is rather good.

TOWER AND MAST EFFECT ON VSWR

Because of sizes and structural differences of different types of towers, investigation of tower effect on the VSWR of a side-mounted antenna is a difficult task. The mutual impedance of a vertical antenna increases as its spacing to a tower decreases. Fortunately for the case where the spacing is less than 0.1λ , these antennas are usually specially designed for side-mounting and are pre-matched with a tower of average size or a mast of fixed diameter. For the case where the spacing is greater than $\lambda/4$, the VSWR effect is likely small. At least, it is quite safe to make such assumption in the case where the diameter of a tower or mast is small compared with wavelength. Following is a good example to this effect.

A special 160-MC colinear array with gain of 4 db over dipole has been de-

signed for a radio system to give coverage flexibilities such as shown in Figures (1) and (2) by using a parasitic reflector. This was done by using a long 2-inch pipe parallel to the array at spacings from 18 to 36 inches. It is seen from Fig. 9 that in the 160-MC region, the VSWR effects of this pipe at 18 inches away from the array to give a cardioidal pattern and 36 inches away to give bi-directional pattern, are not very great.

ACKNOWLEDGEMENT

The authors are thankful to Mr. M. W. Scheldorf of Armour Research Foundation,

formerly of ANDREW Corporation. In the preparation of this paper, numerous references have been made to several of his ANDREW Technical Reports.

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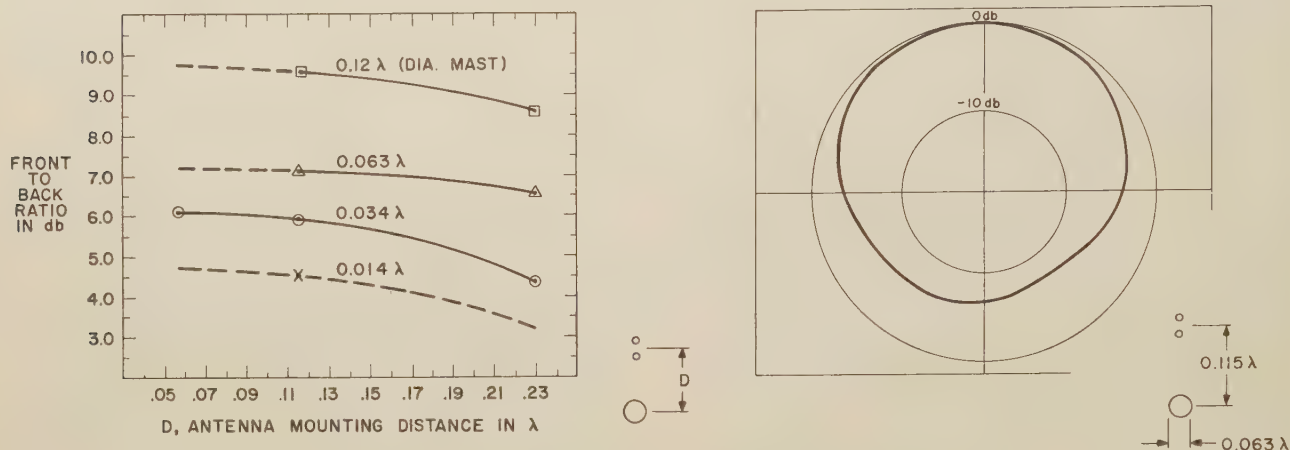


Fig. 1. Front-to-back ratio of a dipole near long mast.

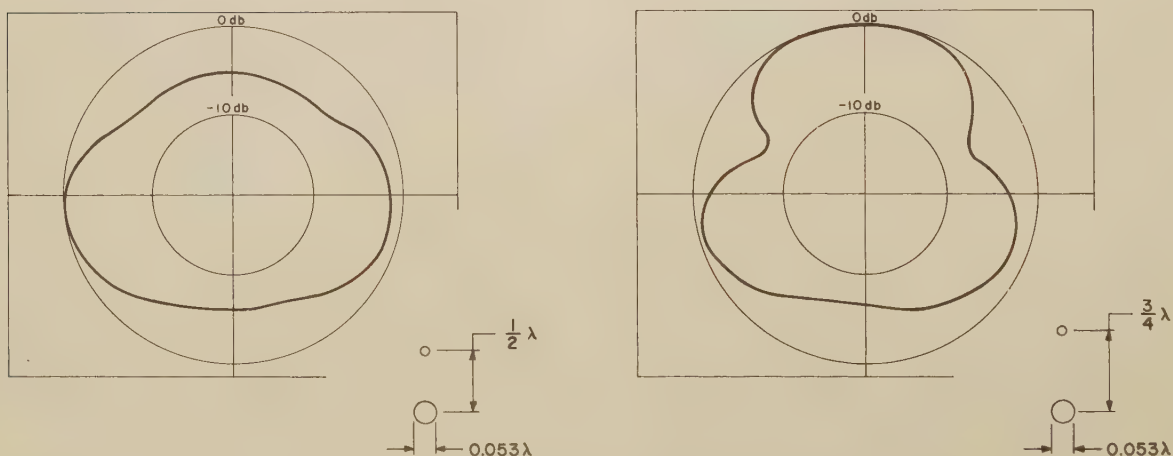


Fig. 2. Patterns of dipole with large spacing to mast.

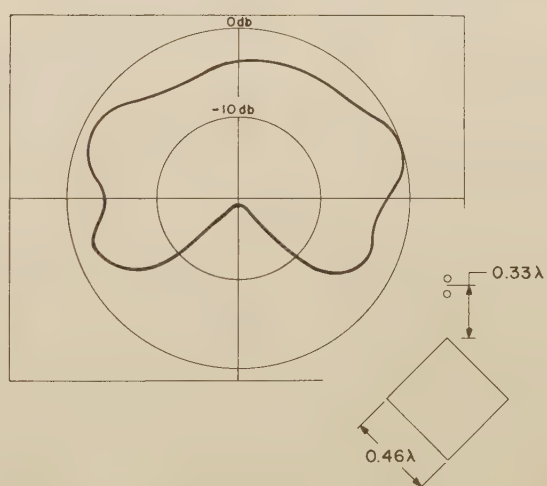
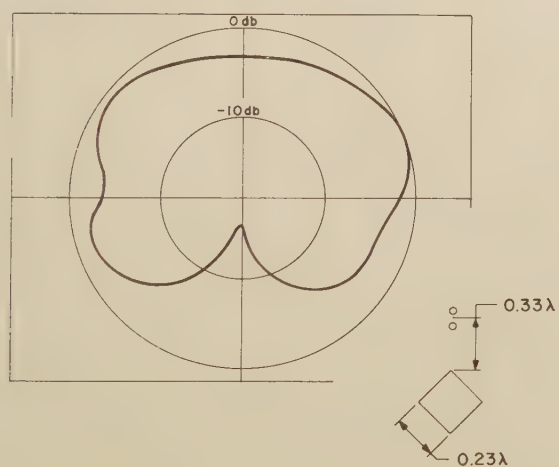
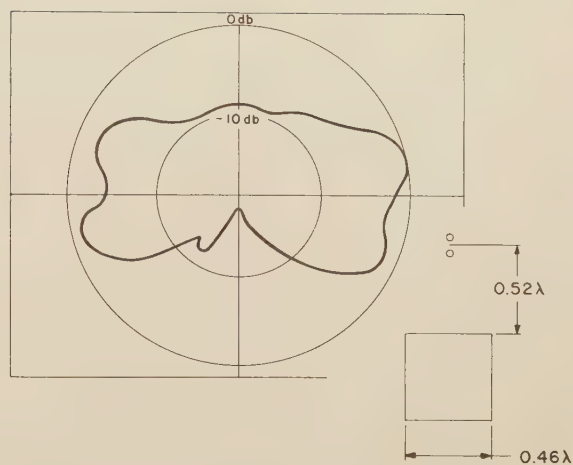
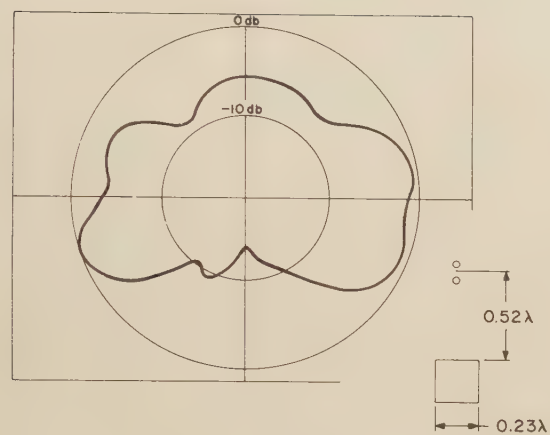


Fig. 3. Effects of large towers on dipole pattern.

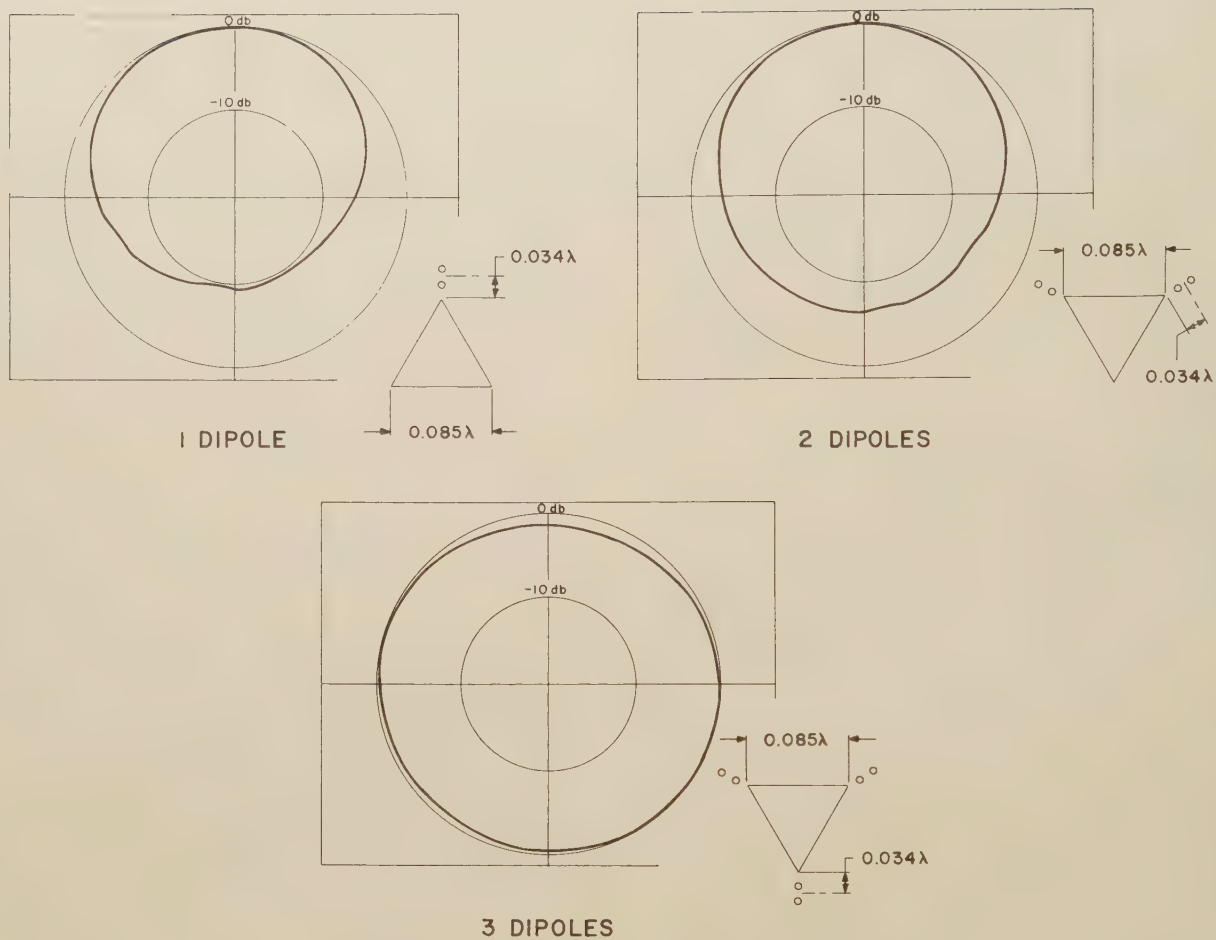
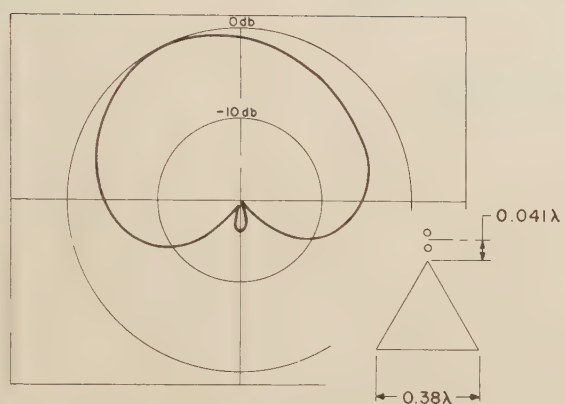
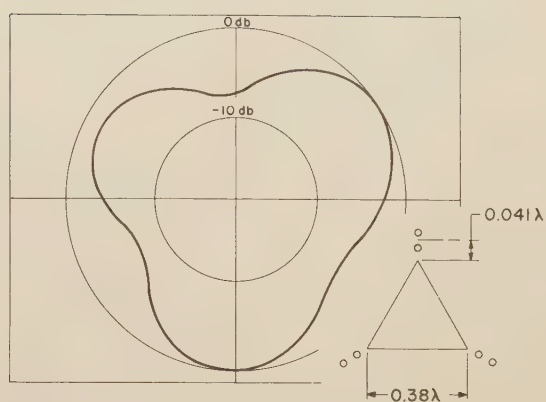


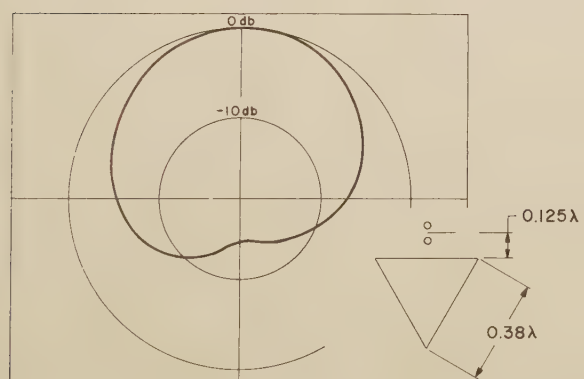
Fig. 4. Effect of tri-angular tower on dipole pattern and methods for circularity improvement.



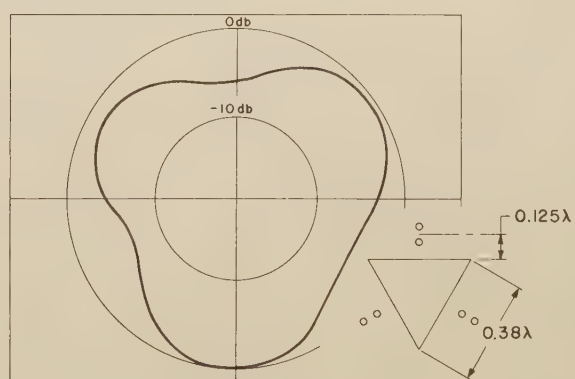
F/B RATIO = 15 db



CIRCULARITY = ± 4 db



F/B RATIO = 14 db



CIRCULARITY = ± 3.5 db

Fig. 5. Effect of medium size tri-angular tower on dipole pattern.

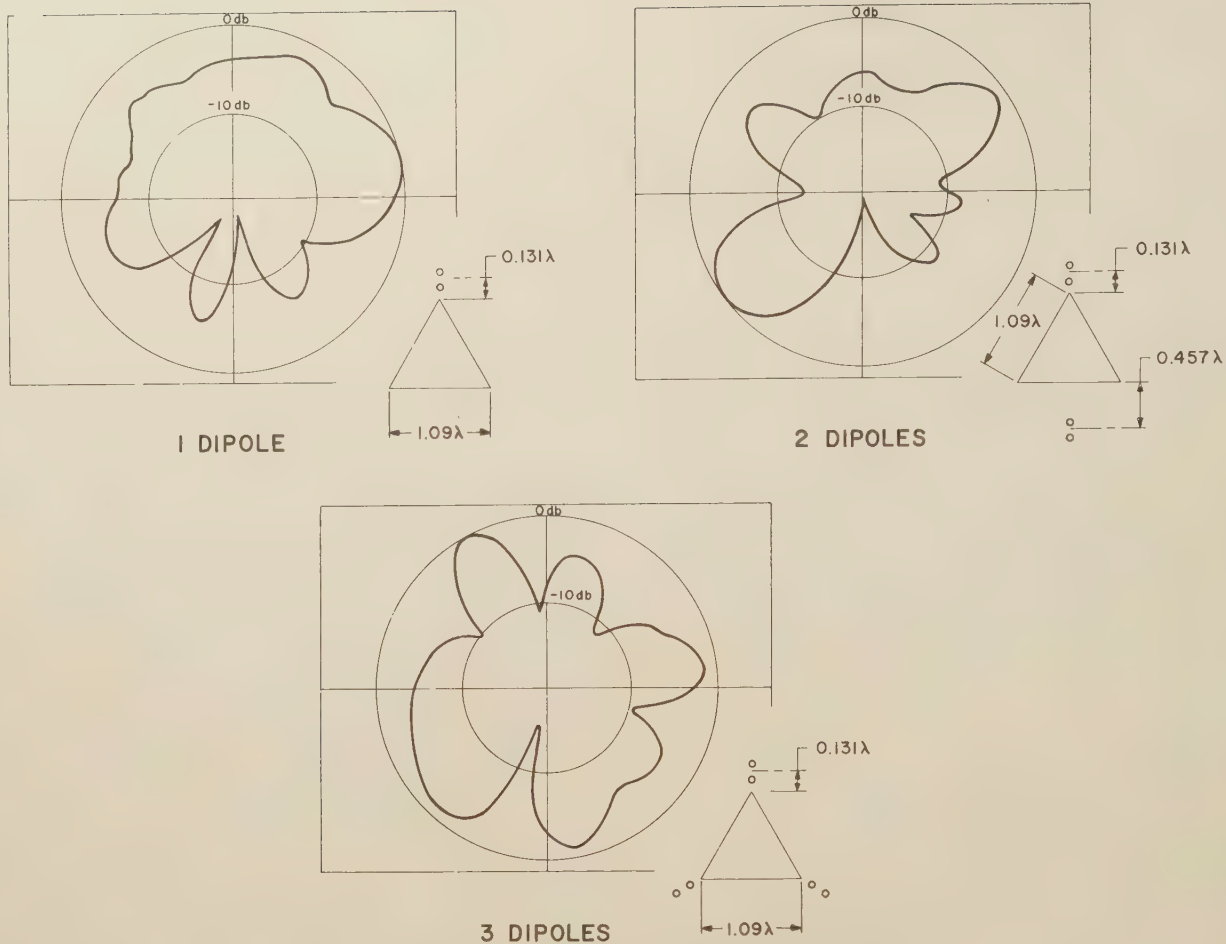


Fig. 6. Effect of extremely large tower on dipole pattern.

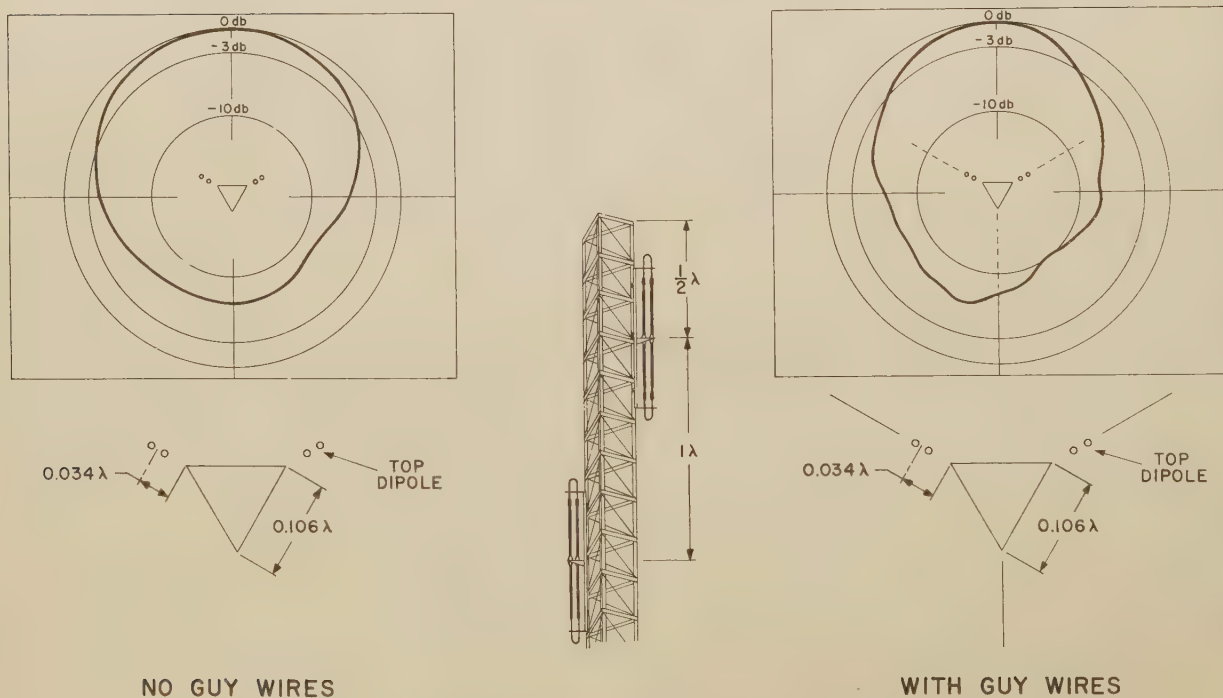


Fig. 7. Effect of guy wires on the pattern of a 2-bay dipole array.

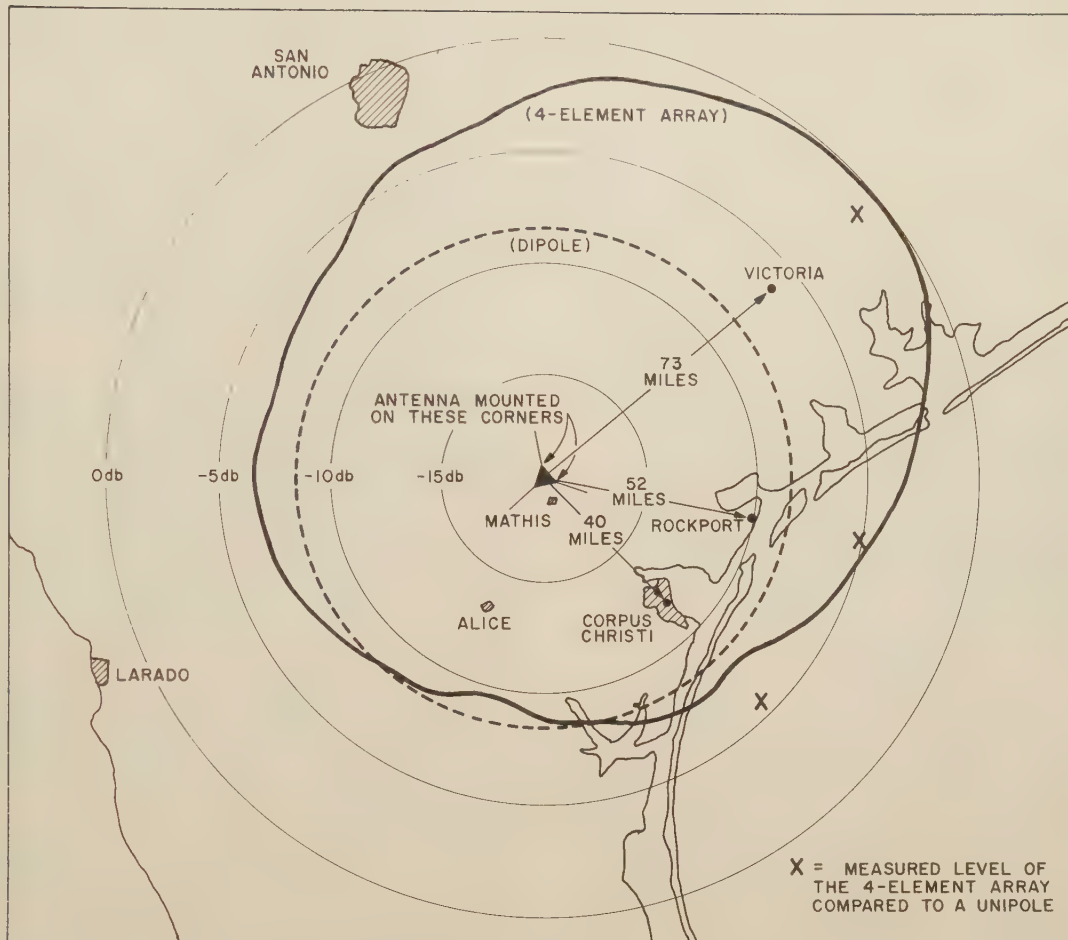


Fig. 8. Field measurement of a 4-bay side-mounted array.

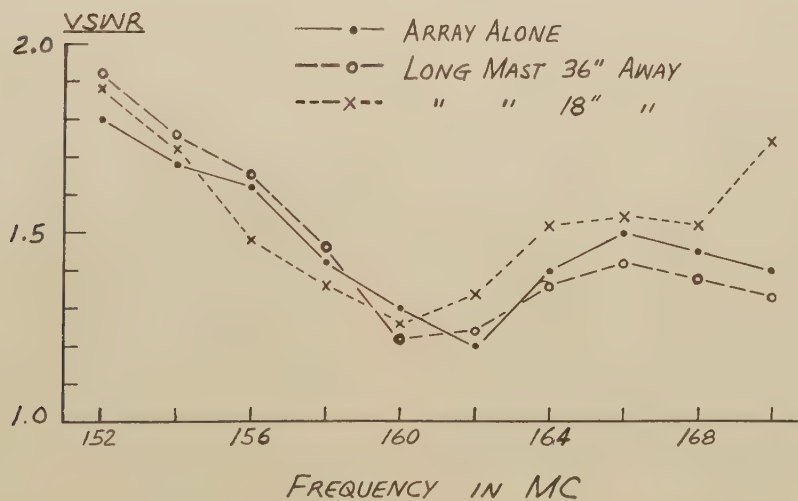


Fig. 9. VSWR effects of a long 2-inch pipe on a 4-db colinear array.

SYSTEM PERFORMANCE, COMPATIBILITY, AND STANDARDS

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Summary

There are two prime purposes of Industry Standards for 2-Way Land Mobile Radio Equipment. They are: 1.) To assure high standards of customer performance and 2.) to assure system compatibility. Examples are given of standards written for both purposes. For example, the logic behind receiver sensitivity and selectivity standards are discussed. Similarly, the importance of standards which assure compatibility between Tone-Coded Squelch Systems, standards for frequency stability, and standards for interference protection between systems are discussed.

Introduction

Those who take advantage of 2-Way Mobile Radio in the operation of their businesses have two kinds of contact with the equipment. The first is via the manufacturers' specification sheets and the second is through first-hand operating experience in a communication system. If the prospective user is to be assured a high standard of performance from the specifications, he must be assured that what he reads on one specification sheet can be correlated with what he reads on another. If one manufacturer were to specify his equipment with an arbitrary set of data taken by his own method of measurement to satisfy his own definition of terms while other manufacturers were just as arbitrary, the system designer would be hopelessly perplexed. He would know what performance to expect only after making a trial installation. It is the purpose of Industrial Standards, therefore, to establish: 1.) meaningful definitions of specifications, 2.) methods of measurements, and 3.) minimum performance standards. Although a manufacturer is not bound by the standards, if he states that his specifications comply with the applicable Industrial Standard, the customer can be assured a high-level of performance. Exceeding the minimum standards, of course, is part of the challenge of a free-enterprise system.

Definition of terms and methods of measurement but not minimum performance requirements are issued as standards by the Institute of Radio Engineers (IRE), whereas, definitions, methods of measurement and minimum performance are issued by the Electronic Industries Association (EIA).

Equipment Performance Standards

Receiver Sensitivity

The most important receiver specification affecting the use of 2-Way Radio is sensitivity. Two methods of measuring sensitivity are recognized by the Electronics Industries Association (EIA) Standard RS-204. The oldest and probably the most used is the quieting method. This procedure measures the on-frequency signal level required to reduce the noise output of a receiver by 20 db. The quieting sensitivity measurement is limited to receivers using one or more limiters and a discriminator. Further restrictions should be placed on this method due to audio response and receiver IF bandwidth. Both audio response and IF bandwidth may be tailored to improve the sensitivity reading at the expense of receiver audio performance. Due to these limitations quieting sensitivity can not stand alone as a measure of receiver performance.

The second method of measuring sensitivity is by the 12 db SINAD method and is called useable sensitivity. This is a measurement of Signal + Noise + Distortion to Noise + Distortion Ratio as produced by a standard modulated signal. The result is a measure of the input signal level which, when modulated by standard modulation, (3.3 kc deviation for narrow-band at 1000 cps) will produce a 12 db SINAD. Figure #1 shows four typical constant SINAD curves for a low-band NB receiver. The rated sensitivity of this receiver is indicated by the arrow at a deviation of 3.3 kc. The constant SINAD curves give a complete picture of a receiver's sensitivity whereas the useable sensitivity is a single point on the 12 db curve. An analysis of these curves will give an indication of the receiver noise bandwidth and its limiting characteristics. The closeness of the curves is an inherent characteristic of FM and the limiter performance of a particular receiver. The bending of each curve above the 5 kc deviation level is controlled by the receiver noise bandwidth. The noise bandwidth of the receiver shown on Figure #1 is slightly greater than 14 kc as indicated by the input at 7 kc deviation, .25 μ v as compared to the input at 3.3 kc, .16 μ v to produce a 12 db SINAD.

Receiver Selectivity

The next most important receiver specification, in the minds of many users, is the selec-

tivity of a receiver. The selectivity of a receiver as defined by the Institute of Radio Engineers in Standard 52IRE17.51 is "that characteristic which determines the extent to which the receiver is capable of differentiating between the desired signal and disturbances of other frequencies". A similar definition appears in EIA Standard RS 204. In both cases the definition and measurement procedure requires the use of at least two signal generators. Although the procedure for the measurement of selectivity was established over 10 years ago, there are a few equipment specifications listing selectivity values which are obviously measured by an obsolete method. The number values appearing opposite the selectivity row in equipment specifications may be recognized by attenuation figures of 100 db or greater at frequency spacings of 15 to 30 kc. This obsolete method, not always specified by name, is the 20 db quieting method. The so-called 20 db quieting method of selectivity measurement is a single signal procedure which not only lacks professional recognition but which actually produces misleading results. For these reasons alone, it should not be used in manufacturer's or purchaser's specifications as a criteria for comparing receiver selectivity performance.

Measurements of 20 db quieting characteristics need not be completely eliminated from equipment specifications for they do serve a useful purpose and may eventually be recognized by either EIA or IRE Standards but not as a measurement of selectivity. The 20 db quieting characteristics have some similarity to the selectivity characteristics of a receiver but unfortunately it departs from selectivity in two areas. These two areas are of major importance being at the nose and high up on the skirt. The most serious inaccuracy occurs beyond about the 70 db level where the error can be from zero to infinity depending on the particular point at which measurement is taken.

Figure #2, a complete 20 db quieting "selectivity" curve, is a good illustration of the region of greatest error. Also shown on Figure #2 is a two signal selectivity curve of the same receiver measured by the EIA standard method. Observe that the 20 db quieting curve has been extended beyond 160 db on both sides. At this point it was necessary to stop the measurement procedure due to overheating of the antenna transformer. It is estimated that this quieting curve would extend upward indefinitely, indicating a selectivity approaching ∞ at +15.5 K and -15 kc. By comparing the 20 db quieting curve to the two-signal selectivity curve on obvious similarity can be seen in certain regions. These regions of similarity are dependent on a particular receiver design. Therefore, it would be ill-advised to express selectivity by 20 db quieting measurements for any frequency off the channel frequency.

The two-signal selectivity of the receiver as shown in Figure #2 is about 50 db at +15 kc

and increases to 88 db for off-channel frequencies of +35 kc. To express this receiver's selectivity by any other values, such as, 100 db at +15 kc is an obvious deviation from engineering standards. Furthermore, the use of 20 db quieting "selectivity" figures to evaluate system performance in the presence of interference will lead to considerable error. In the case shown, the error would be about 50 db at 15 kc off the channel frequency. Thus, a system designed on the basis of 100 db selectivity at 15 kc would be degraded by a signal equivalent to 50 db or 10,000 times stronger than the weakest useable desired signal.

The 20 db quieting curve, although not a measure of receiver selectivity, does provide certain useful information. Figure #3 introduces a new curve, squelch opening, which is essentially identical to the 20 db quieting curve introduced in Figure #2 and duplicated in Figure #3. This squelch opening curve was taken by adjusting the receiver's squelch for "standard test" conditions. Standard test conditions for squelch opening is arbitrarily set at 3 db below the SINAD sensitivity which is about 5 db below the 20 db quieting sensitivity. Then the squelch opening curve has a reference which is 5 db below the 20 db quieting curve reference. If the squelch opening curve were taken by first adjusting the squelch to open at the 20 db quieting level, the two curves would be identical, thus the 20 db quieting "selectivity" is basically a squelch opening characteristic. The $\pm$ 20 db quieting "selectivity" if properly named "squelch opening" would provide a useful specification in system performance evaluation.

Vehicular Noise Interference

Vehicular noise radiation is a major source of interference which degrades the performance of two-way radio. There is no EIA Standard as yet for measuring the degradation of performance produced by vehicular noise; however, the need is well recognized. As a first step in establishing a standard, EIA has participated jointly with the Society of Automotive Engineers, (SAE), in a program of Standardizing Methods of Measurement of Vehicular Noise Radiation. As shown in Figure #4, SAE has established methods of measurements and limits which are primarily suited to reducing interference to TV receivers. These measurements are taken at a distance of 50 feet and are expressed in units of uv/meter/kc. After studying the methods of measurement proposed by the SAE, IRE Sub-Committee 27.7 decided that measurements made at 50 feet were not appropriate for evaluating interference created to receiving equipment installed in the vehicle. The IRE Sub-Committee 27.7 has proposed a new standard of measurement which is now in the approval process. This standard utilizes a broad-band antenna located four feet from each bumper thus providing two readings for each frequency measured. The average noise-level curve of Figure #4 gives typical values of

radiation levels measured by the IRE method for vehicles meeting but not exceeding the SAE limits. Levels of interference measured by the IRE method are expressed in db above 1 $\mu\text{v}/\text{mc}$. A typical highband reading would be 94 db above 1 $\mu\text{v}/\text{mc}$ which is equivalent to 54 db above 1 μv as received by a narrowband receiver.

Vehicular noise is primarily of the impulse type; thus, the 54 db above 1 μv is a peak voltage level of short duration. If the total power represented by a peak impulse of 54 db above 1 μv were received by an installed receiver, the degradation of receiver sensitivity would be about 24 db as shown by the upper limit. The actual degradation will be between the upper limit and the lower limit depending on the condition established in a particular vehicle. The variation of about 20 db between the upper limit and the lower limit is in part caused by the radio frequency characteristic of the vehicle and its installed antenna. The radio-frequency characteristics of a vehicle creates points of minimum and maximum signal for each radio frequency and are dependent on the location of the antenna. Points of minimum and maximum signal can be moved about by changing the physical parts of the vehicle including noise suppression devices. Degradation limits of lowband receiver have about the same range but at higher level.

System Compatibility Standards

Being assured of high standards of performance would be of little value to a peruser of equipment specification sheets if the equipment were not designed to be compatible within and between systems of units. For example, the finest AM receiver built would be worthless for voice communication in an FM system. This, of course, is an extreme case which illustrates the necessity for standards which will assure compatibility.

Continuous Tone-Coded Squelch Systems

In recent years, the crowded-channel conditions have lead to Continuous Tone-Coded Squelch Systems (CTCSS) in order to reduce botherance from other systems on the same RF channel. Difficult as it now is to be sure that tone-codes are assigned in accordance with Standards prepared by the Electronics Industry Association, there would have been utter chaos had no standards been set.

The fundamental CTCSS standard is that of the tone-code frequencies to be used because it is the choice of these frequencies that determines whether units in a given system can communicate with each other and whether they will interfere with other units not in the same system.

There is a choice to be made in the selection of tone-code frequencies. Where in the audio band should they lie? Let us suppose that we were to use a tone frequency of 3000 cps and

the Q of the tone decoder were 50. The bandwidth would then be $BW = \frac{f_o}{Q} = \frac{3000}{50} = 60 \text{ cps}$. If

a low-frequency tone of 100 cps were used, the bandwidth would then be 2 cps. Since falsing due to noise is an extremely important consideration in CTCSS system, it is important to note that the noise-power increase in the high-frequency vs. the low-frequency case is $\frac{60}{2} = 30 = 15 \text{ db}$. Thus, the low-frequency tone

system operates at a 15 db signal-to-noise advantage over the high-frequency system, assuming constant Q. It is also important to consider which tone one is least likely to hear at the receiver output. Figure #5 shows the response of a typical receiver audio system with consideration of the speaker and ear response. The heavy curve indicates that the actual perceived level at 100 cps is 37 db lower than the level at 2000 cps. This means that the non-tone-squelch equipped user would hear less of the 100 cps modulation than he would of 2000 cps for the same deviation.

With all things considered, then, the low frequencies are less of a nuisance to non-equipped users, are easier to keep out of the audio, and operate with a better signal-to-noise ratio. The absolute tone frequency is a function of many things, however. For instance, the time which is required for the tone decoders to operate is a function of circuit Q, or more specifically, the operating bandwidth. This bandwidth, along with the guard bands, determines the tone-channel spacing - a compromise, of course, between speed of response and the number of tone codes required to satisfy the crowding situation.

The guard bands are determined both by the magnitude of RF carrier deviation produced by the tone and the tone-frequency stability in order to prevent interference between adjacent tone-channels. All of these factors add up to tone-code frequencies being standardized at approximately 3.5% spacing in the 67 cps to 254 cps range. The RF carrier deviation is restricted to + 0.9 to 2.0 kc in wide band systems and + 0.5 to 1.0 kc in narrow band systems. The frequency stability must be within + 0.5% of the assigned standard frequency over the extremes of environmental conditions.

It is possible - indeed easier - to design tone-coded squelch systems in a fashion where severe discourtesy and disruption may result. Consider the case of the operator of a tone-coded unit in his car. When he hears nothing from his loudspeaker, it means that his base station is not transmitting. But another base station on the same RF frequency may be on the air! If our mobile operator simply lifts his microphone to talk, he may very well break in on a conversation already underway on the channel. It is important, therefore, that courtesy be the watchword. This means that the tone-coded operator must disable his tone-coding in order to monitor the channel

before transmitting. Human beings being what they are, it is wise for the manufacturer of equipment to perform this monitor function for the operator in such a way that he does not have to do anything but prepare to transmit. One favored method is to un-mute the receiver when the microphone is lifted from the hangar.

RF Carrier Frequency Stability

The subject of channel spacing and the attendant subjects of RF carrier stability, receiver noise width, skirt selectivity, desensitization, intermodulation and impulse noise performance cannot be treated fairly nor in detail in a paper such as this - yet they are of the utmost importance in assuring system compatibility.

Channel spacings are made by the FCC, although industrial groups are frequently called upon for technical advice before rule makings are proposed or finalized. Spacings tend to be evolutionary, however, - each new spacing tending to split the preceeding set. For example, the high band channels (150-174 mc) have been split from 120 kc to 60 kc and now to 30 kc. Similarly, in low band (25 - 50 mc) there have been the splits from 80 kc to 40 kc to 20 kc.

Figure #6 shows the average interference range to be expected as a function of channel spacing with typical present-day equipment. It is clear that operation on 20 kc channels is on a much steeper part of the interference curve than at 30 kc. We might conclude that the low-band spacing of 20 kc was not quite so wise as the 30 kc high-band split. Yet there was the very large factor of economics involved when that decision was made. Had anything been done except split the 40 kc channels, even then-existing systems would have had the rather sizeable expense of changing frequency. There were also the economics of available spectrum to consider. Splitting to 20 kc channels yielded more channels than some other wider spacing.

Figure #7 is a summary of the receiver nose-selectivity situation as it exists today. This curve illustrates that the nose must be wide enough to accept the modulation even during the condition of maximum transmitter and receiver drift. The allowable magnitude of this drift is determined by two sets of boundary conditions. The first we have seen in Figure #6 as the effect of spacing on interference. The other limiting condition is the receiver performance in the presence of impulse noise. The further the desired frequency is from the center of the pass band, the more severe will be the phase shift of the noise pulse as it traverses the frequency-shaping system. It is this phase response that determines, to a major degree, the percentage of receiver noise width that the signal may be allowed to drift. Thus the reasons for Industry

Standards groups, such as the EIA and IRE, having established standards on receiver selectivity, and transmitter and receiver stability. It is this type of analysis that points out the need for a standard of impulse noise performance along with a maximum level of impulse noise that may be radiated by vehicles.

Interference Protection

Not only must systems be compatible within themselves, but they must have a degree of interference protection from adjacent channel and other systems. A proposed EIA standard for Land-Mobile Systems outlines Standards and Methods of Measurement for 1.) System Interference Rejection, 2.) System Intermodulation Interference Rejection, and 3.) System Undesired Signal Rejection. Measurement procedures for these standards are shown in Figure #8 - a, b, and c, respectively.

The System Interference Rejection measurement of Figure #8-a is a measure of the ability of a system to prevent greater than a 6 db reduction in the 12 db SINAD when a single undesired signal is introduced into the system.

The measurement of Intermodulation Interference Rejection of Figure #8-b is a measure of the ability of a system to prevent greater than a 6 db reduction in a 12 db SINAD when two or more undesired signals are fed into the system.

The measurement of System Undesired Signal Rejection of Figure #8-c is similar to the measurement of System Interference Rejection except that it measures the ability of a system to prevent audible or functional outputs due to external signals in the absence of the desired signal. The squelch is adjusted to a prescribed critical point for this measurement.

Conclusion

The foregoing are examples of Industrial Standards that have been written to assure high levels of performance and system compatibility when equipments are designed to comply with them.

It is important, however, to know the meaning of the Standards behind the performance specifications and to be sure that manufacturers do comply with them. It would be inexcusable, for example, to substitute a pseudo specification for one which is recognized as an Industry Standard.

These instruments then become valuable both to the manufacturers, who work to common standards, and to discerning users who can select equipment knowing that it was designed to meet certain minimum standards.

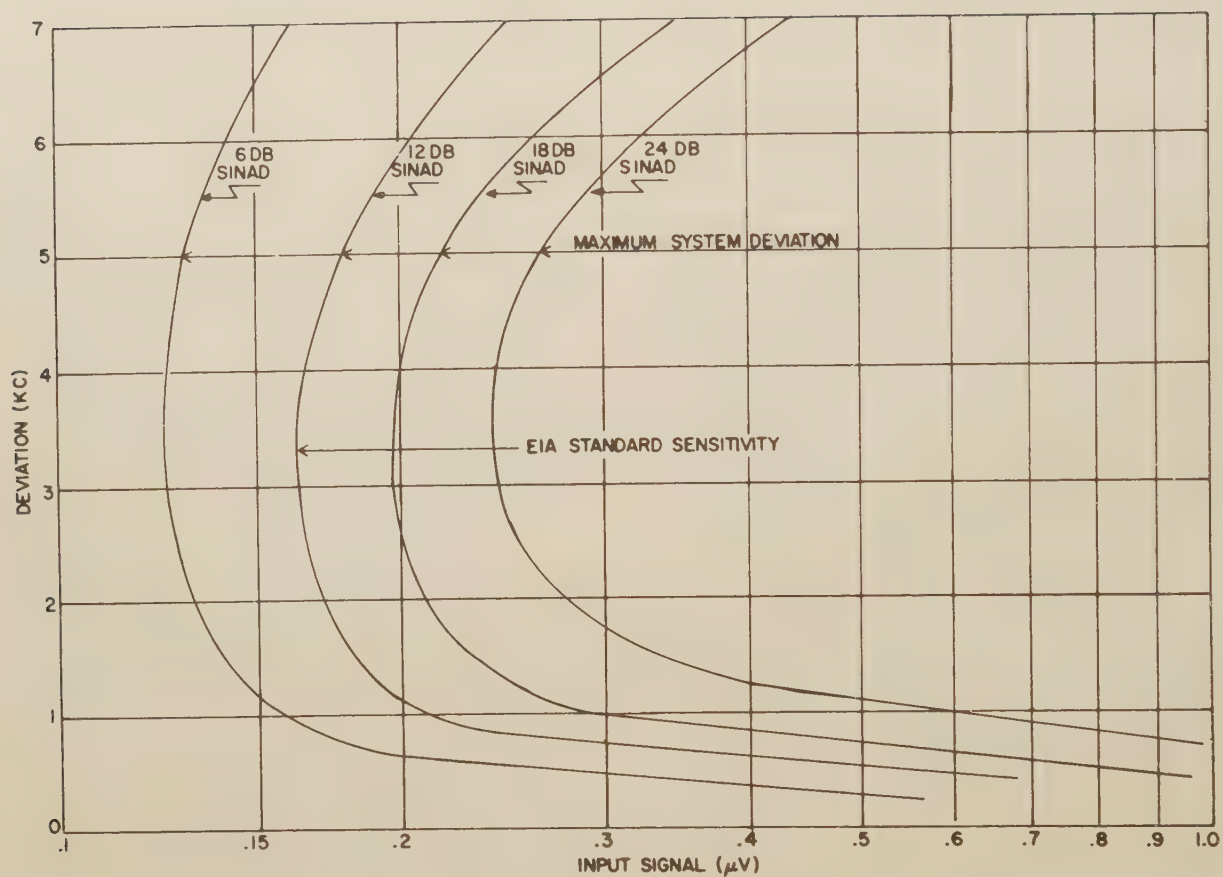


Fig. 1. Typical low band NB receiver sensitivity (constant sinad ratio).

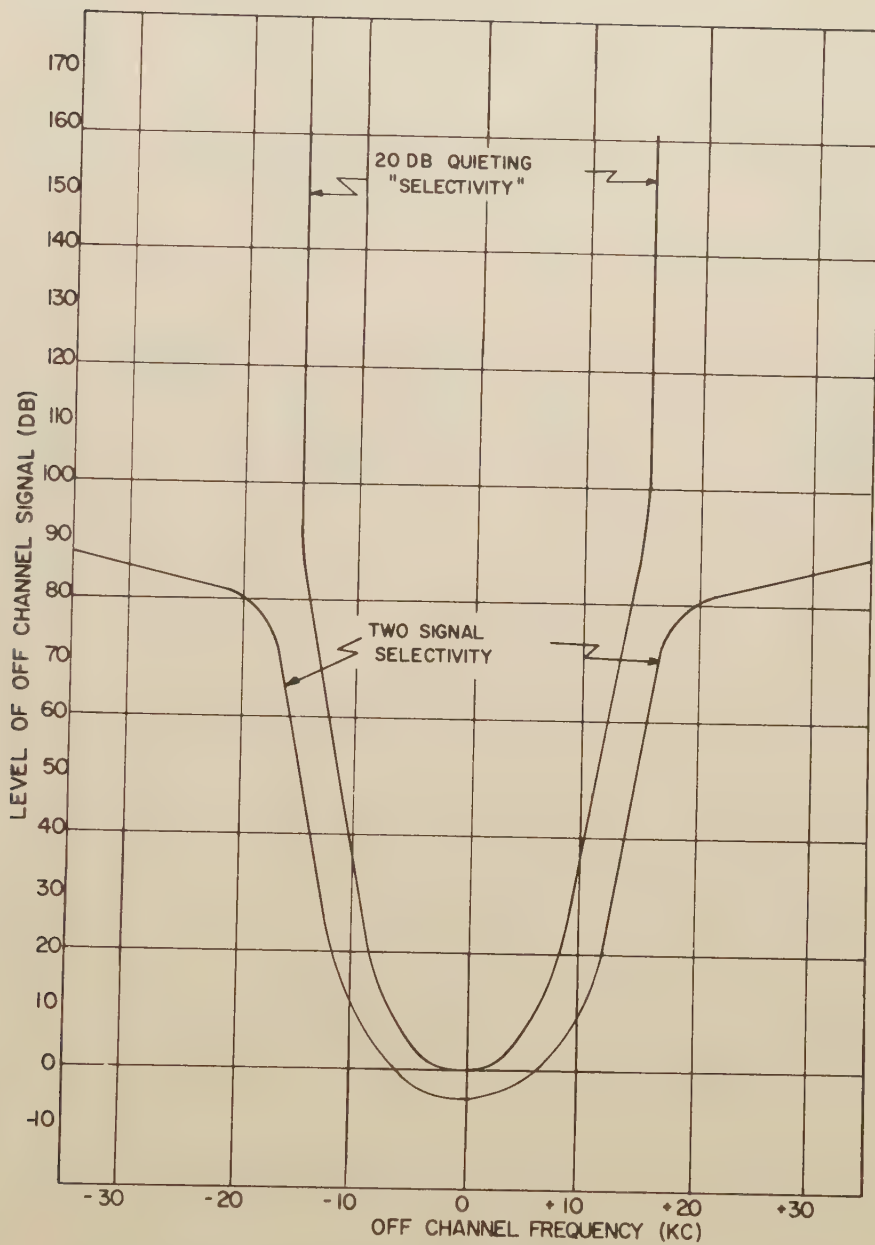


Fig. 2. Typical low band narrow band receiver characteristics.
20-db quieting vs two-signal selectivity.

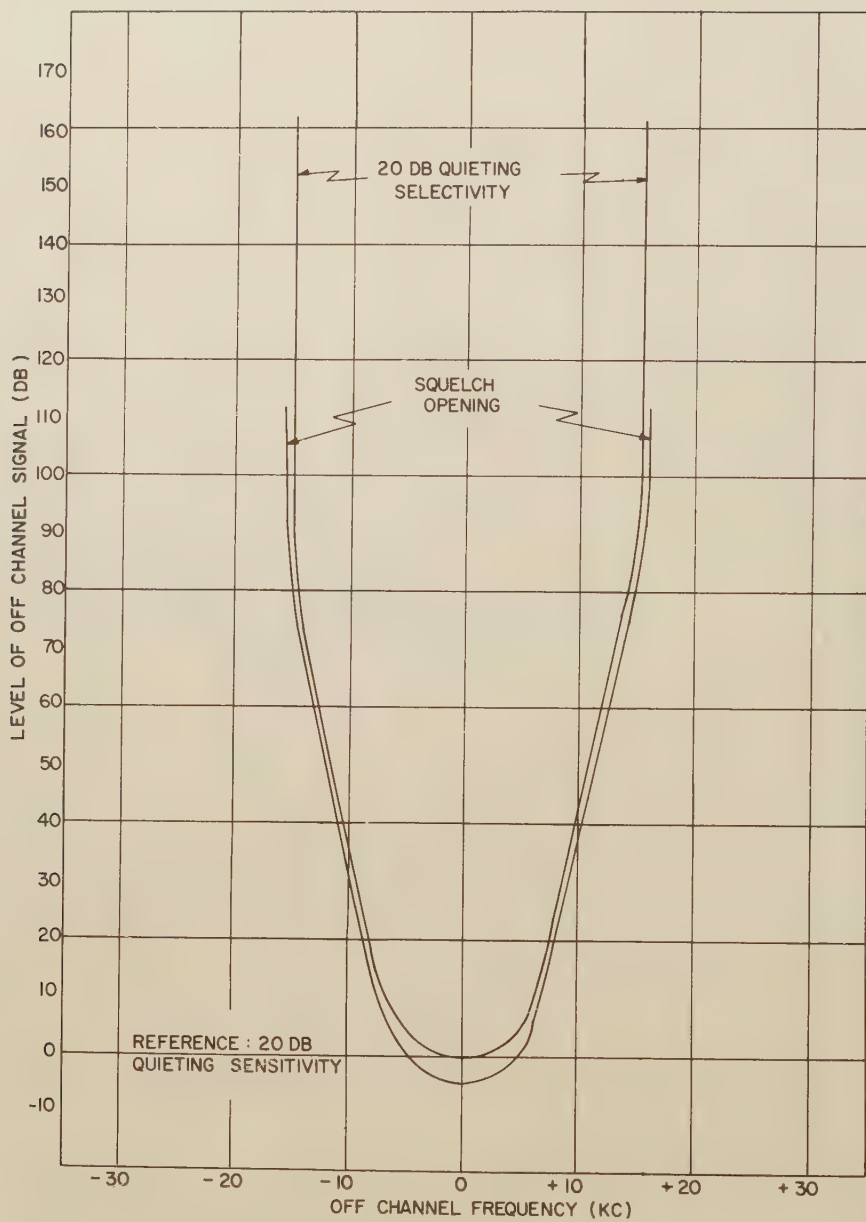


Fig. 3. Typical low band narrow band receiver characteristics.
20-db quieting vs squelch opening.

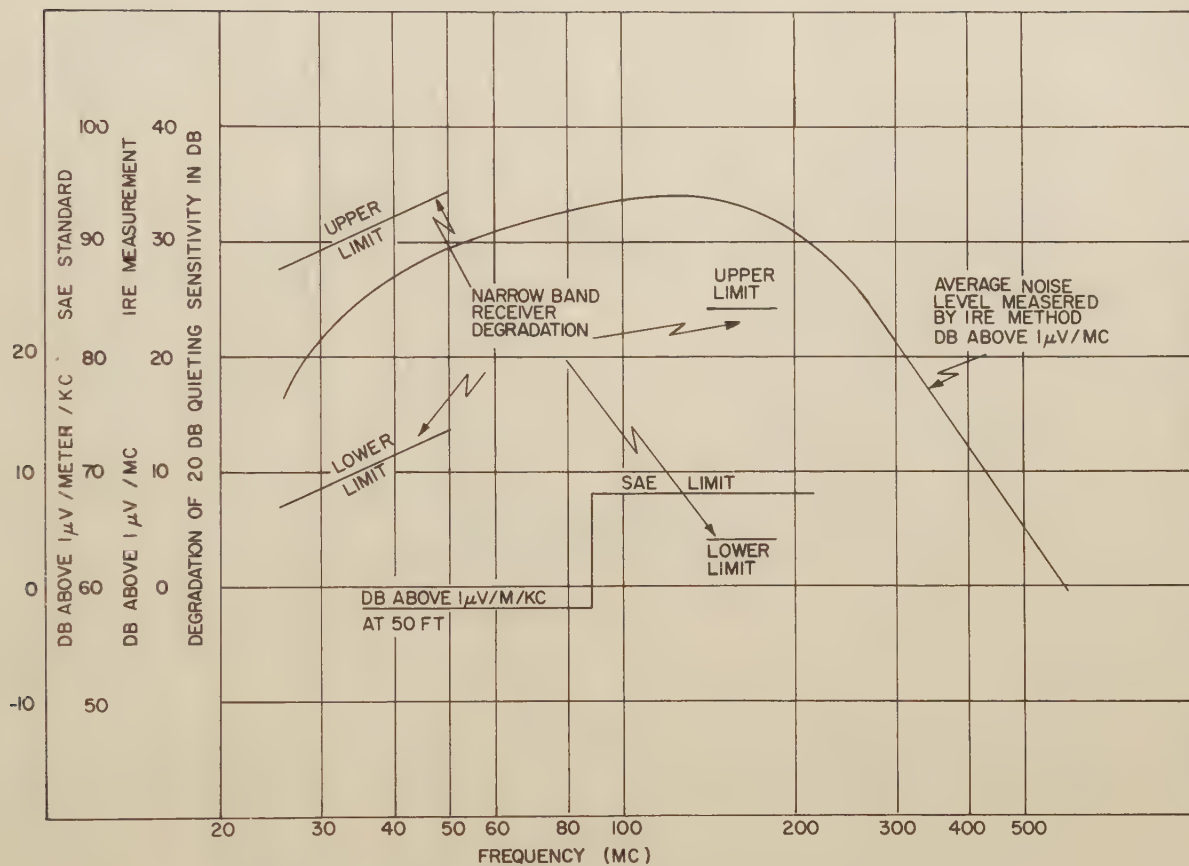


Fig. 4. Vehicular noise radiation SAE limits, IRE measurements and receiver degradation.

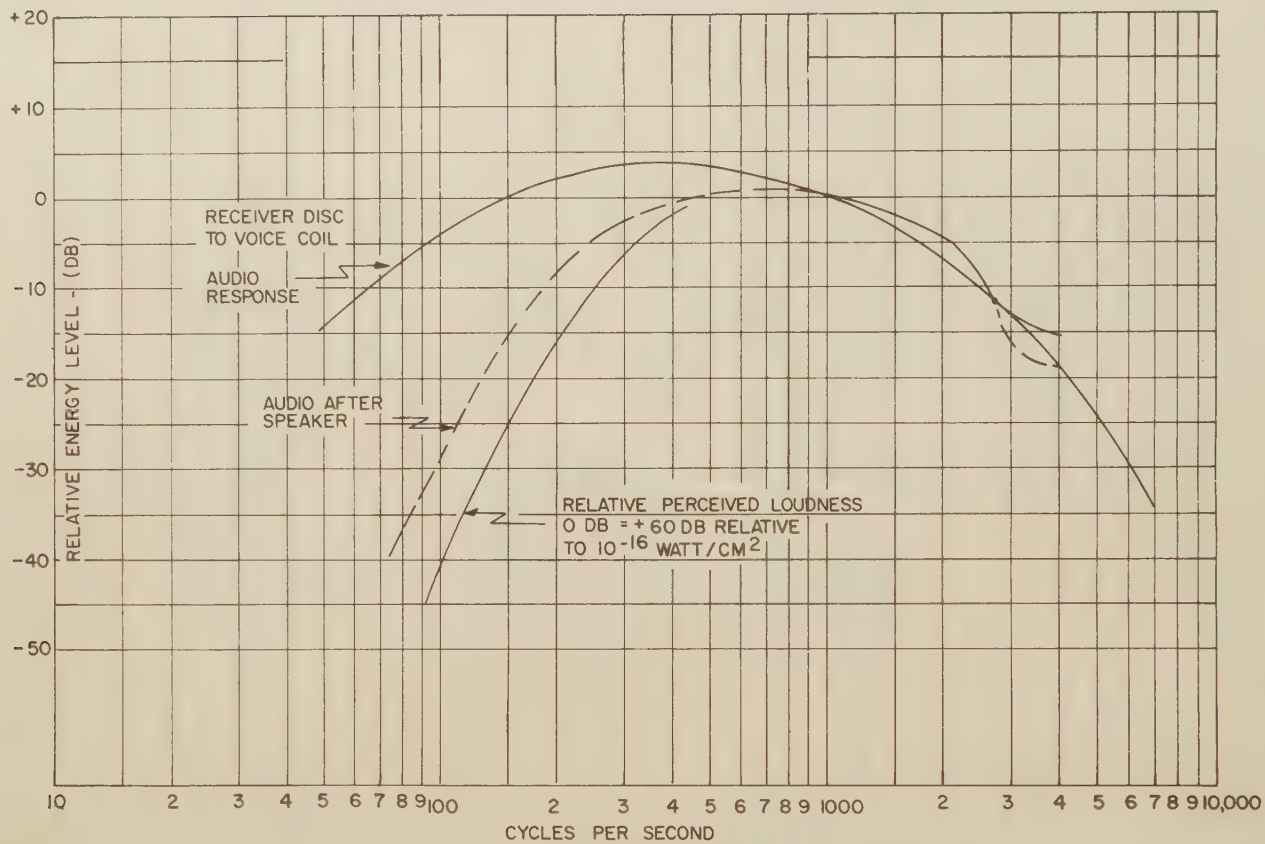


Fig. 5. Receiver response from discriminator to listener's ear.

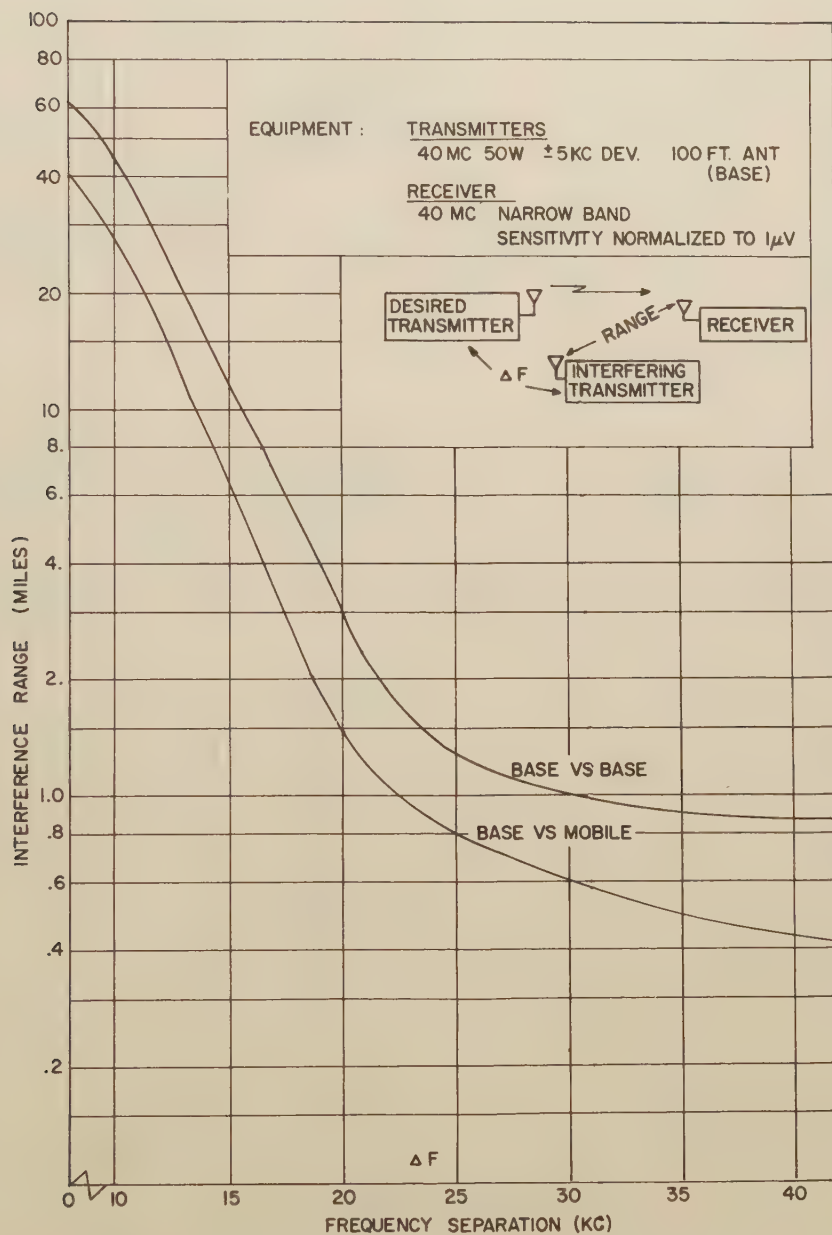
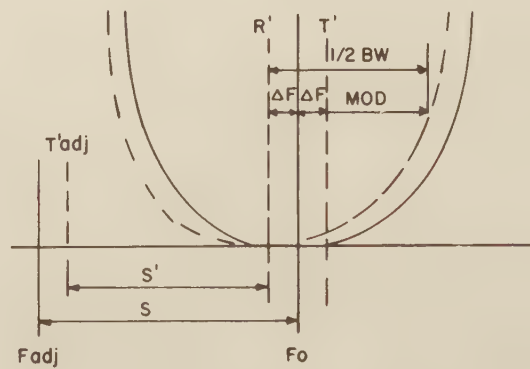


Fig. 6. Average interference ranges (for 6-db reduction in 12-db sinad).



Fo = DESIRED CHANNEL

Fadj = ADJACENT CHANNEL

ΔF = MAX. FREQ. DRIFT

R' = REC. FREQ AT MAX. DRIFT.

T' = TRANS FREQ. AT MAX. DRIFT.

S = CHANNEL SEPARATION

S' = CHANNEL SEPARATION AFTER DRIFTS

FREQ. (MC)	S (KC)	STAB. (PPM)	ΔF (KC)	$2 \Delta F$ (KC)	S' (KC)	MOD. (KC)	1/2 BW (KC)
25	20	2	.5	1.0	19.0	± 5	6.0
50	20	2	1.0	2.0	18.0	± 5	7.0
150	30	5	.45	.9	29.1	± 5	5.9
175	30	5	.875	1.65	28.3	± 5	6.65
450	50	5	2.25	4.5	44.5	± 15	19.5
470	50	5	2.35	4.7	44.3	± 15	19.7

Fig. 7. Receiver noise selectivity.

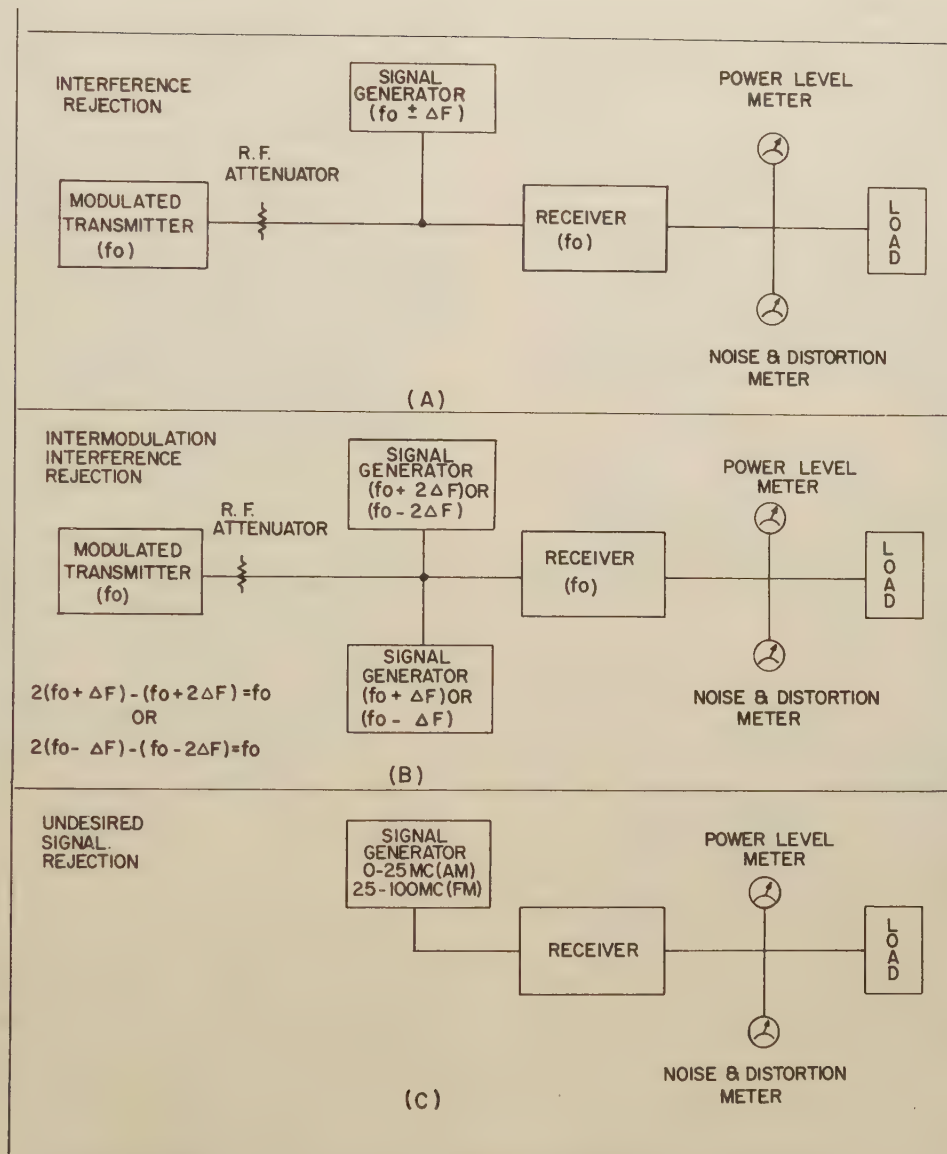


Fig. 8. System interference performance measurements.

A PERSONAL TWO-WAY VHF RADIO COMMUNICATION SYSTEM FEATURING MODULAR CONSTRUCTION

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Summary. This paper describes a VHF "personal" communications receiver assembled entirely from individual circuit modules, which are constructed with standard, commercially available, subminiature components. Defective modules can be rapidly replaced, and later repaired or modified. Other receiver features include 1 microvolt sensitivity, double conversion, a crystal filter selectivity package at the high I. F. frequency, and a novel squelch gating circuit.

The companion transmitter, more conventional in construction, also is subminiaturized. It features a 1 watt power output, a transistorized modulator, and a DC-DC converter type of power supply with rechargeable nickel cadmium batteries. The shirt pocket size receiver and the transmitter occupy a combined volume of less than 85 cubic inches and together weigh less than 4-1/2 pounds.

Introduction

The modular concept of packaging electronic equipment is becoming increasingly common in new equipment designs. Ease of repair by first echelon maintenance crews is largely responsible for its success in military design, since modules can be replaced at the operational level and repaired later at the base or depot level.

Subminiature equipment of modular design has tended to emphasize throw-away modules. Although this can be costly, little else can be done in the case of "grown" or plated modules since individual circuit components are either non-replaceable or non-existent. Where extreme miniaturization is not the determining factor, repairable modules, constructed from standard subminiature components, provide all of the advantages of modular construction and retain those of standard wiring construction. From a production standpoint, repairable modules are advantageous because: 1) they can be constructed by standard or slightly modified standard production methods; 2) design changes during production can easily be incorporated and completed modules can be reworked with a minimum of scrap; and 3) defective modules can be repaired on the production line.

Receiver Construction

A "personal" size FM communications receiver, incorporating the modular techniques described, has been constructed at Bendix Radio. Figure 1 is an overall view of the receiver, which operates on any channel in the frequency band of 148-175 mc. The package occupies a volume of 23.7 cubic inches and weighs 1 pound. Its modules account for only 5.6 cubic inches of this volume; the balance of the space being used by the batteries, antenna, controls and loudspeaker. An external earphone may be used in lieu of the internal speaker.

This is not the smallest receiver of its kind ever built and no attempt was made to make it so. The unit is small enough, however, to be carried in a shirt pocket, and yet the modules and components are of such a size that they can be maintained by technicians of normal skill, rather than by watchmakers.

Figure 2 shows the receiver with the case removed, revealing the compact, yet accessible, layout.

The requirement that the individual modules be easily manufactured and repaired, yet be of small size, led to the development of a refined "sandwich board" technique of module construction and assembly. This technique is fully described elsewhere¹ and will be outlined here only briefly.

A module is constructed from two printed wiring boards, approximately 1/2" by 5/8" (Fig. 3), and the required subminiature components. The boards, devised by D. J. Houck of Bendix Radio, have V-shaped notches around the edges for component mounting. The notches are designed to retain wire leads of various diameters until the soldering operation can be performed. Interconnections between components are printed onto the boards, and wire leads bring all necessary interconnections out of the modules. A completed module is shown in Figure 4 along with its schematic representation. The module has a volume of 0.27 cu. in., and contains 13 components. This represents an approximate package density of 48 per cubic inch, or 83,200 per cubic foot, less than that achievable with microminiaturization but considerably greater than that achieved with conventional wiring or even flat board printed wiring techniques. Many of the modules used in the receiver have package densities of 90,000 per cu. ft., with some experimental modules going as high as 150,000 per cu. ft.

General Description

The developmental receiver constructed from these modules, shown in block diagram form in Figure 5, is a double conversion superheterodyne of conventional design. It contains a tuned RF stage and a crystal filter selectivity package located in the 10.7 MC 1st IF amplifier chain. This arrangement keeps overall receiver stability independent of the 2nd oscillator or 2nd IF amplifier chain, and tends to minimize detuning caused by temperature and transistor aging in this chain. The 2nd IF string is resistance coupled and adjusted to give the effect of being very broadly tuned at 200 KC. Squelch control is accomplished by means of a transistor switch, gated by processed noise. To conserve battery life, all audio stages are cut off when the receiver is squelched.

Twenty-one germanium transistors and four germanium diodes are used in the receiver. Type 3N37 tetrodes are employed in the RF amplifier and first

mixer modules. An NPN type 2N467 is used as a squelch gate. All other circuits use common PNP transistors and diodes such as 2N207 and 2N217, in the audio stages, 2N410 in the low IF, 2N588 in the high IF amplifiers, and 1N116 diodes in the discriminator and squelch modules.

The selectivity package operating at 10.7 MC is a crystal lattice filter, with the sharpness of the selectivity curve dependent on the number of crystals used in the filter. A four crystal network was used yielding a nose-to-skirt ratio (6-60 DB) of 3:1. Eight crystal networks, with ratios of 1.8:1 have been achieved in production runs on other equipments at Bendix and could be incorporated in this receiver.

The squelch module contains a three stage Darlington connected noise amplifier providing 60 DB of amplification, thus increasing the sensitivity of the squelch around threshold. The noise is rectified after amplification and controls a transistor switch which biases the audio amplifiers for normal operation.

For purposes of comparison, a modular block diagram is shown along with the conventional block diagram in Figure 5. Notice that in modular block form certain functions appear which do not show up on conventional diagrams. For example, tuning coils normally are associated with the circuits they tune but in this type of representation they appear separately since because of their size, they occupy an entire module. Similarly all B- filtering and decoupling components are lumped into a single module and thus appear in the modular block diagram.

The receiver is made up of 19 modules of which three are double size. It, therefore, contains the equivalent of 22 single size modules. After assembly, modules can be tested individually to governing specifications to insure interchangeability. Module assemblies are housed in metal cans for increased mechanical strength and electrical shielding. At final assembly the modules are mounted by their external leads to a double sided master circuit board, which also picks up connections for the batteries, speaker, antenna, and controls. The master board is mounted in the upper part of the exterior receiver case (Fig. 2), and the entire assembly is then inserted into the lower portion of the case.

One rather obvious benefit of modular design is the reduction in design effort required to produce variations of a basic design. For example, although this receiver operates between 148-175 MC, the 2nd IF, discriminator, squelch and audio modules also are useable in a receiver operating in another frequency band. Similarly, by changing crystal filter and discriminator modules, the receiver bandwidth may be varied.

Performance

Receiver performance is equivalent to that of units constructed along conventional lines within the environ-

mental limits of the components used, except for the limitations in acoustical power output imposed by the small speaker used in the personal unit. Sensitivity is 1 hard microvolt for a 12 DB SINAD ratio. Audio output is 200 MW to the loudspeaker. Current drain is 15 MA with the receiver squelched and 70 MA at maximum output. Batteries will last for 50 to 75 hours of normal operation.

Transmitter Description

To facilitate field testing of the receiver, a companion transmitter was constructed. This transmitter, shown in Figure 6 along with the receiver, consists of a transistorized, modularized, modulator and subminiature vacuum tube oscillator, multipliers and power amplifier.

No attempt was made at extreme miniaturization in the vacuum tube stages since it was felt that fully transistorized transmitters soon will be feasible from both a technical and economic standpoint. At this writing, transistors are becoming available which will yield required power outputs in the 148-175 MC frequency band.

The transmitter shown occupies a volume of 18.3 cubic inches and weighs 1 pound. Plate power is supplied from a DC-DC converter which is powered from from a bank of rechargeable nickel cadmium batteries. The battery case attaches to the transmitter case adding 39.5 cu. in. to the volume and 2 lb., 5 oz. to the weight.

Transmitter power output is better than 1 watt. Spurious responses are more than 65 DB down. The modulator contains deviation limiting circuitry and is capable of up to ± 15 KC deviation. With a 10% transmit duty cycle, approximately eight hours of operation can be obtained from a set of batteries before recharging is required.

Conclusion

The system described here is a reasonable compromise between extreme subminiaturization and full size equipment. Producibility and maintainability by semi-skilled personnel is still an important economic factor especially in the commercial equipment field. Repairable modules adequately fit this category and at the same time yield designs which are the equivalent, performance-wise, of other more common design techniques.

- ¹"Evaluation of a High Density Approach to Electronics Packaging" - J. Kuller & T. Knight (Paper presented at 1st International Packaging Symposium, Boulder, Colorado, August 1960.)



Fig. 1. VHF receiver which will fit in a shirt pocket.

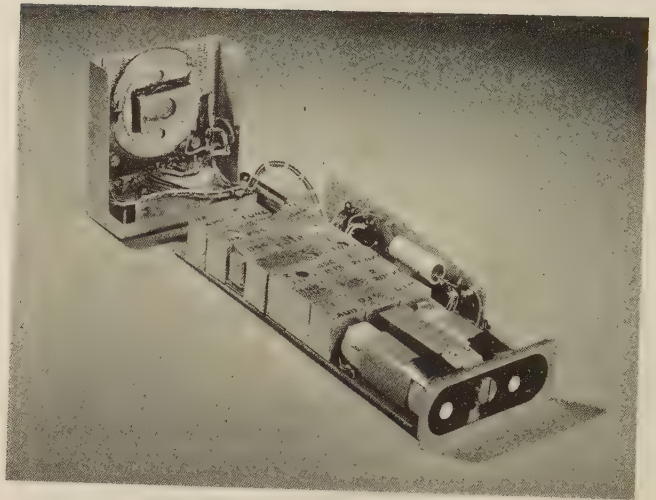


Fig. 2. Receiver is constructed of removable, repairable modules.



Fig. 3. Module components are mounted between notched printed wiring boards.

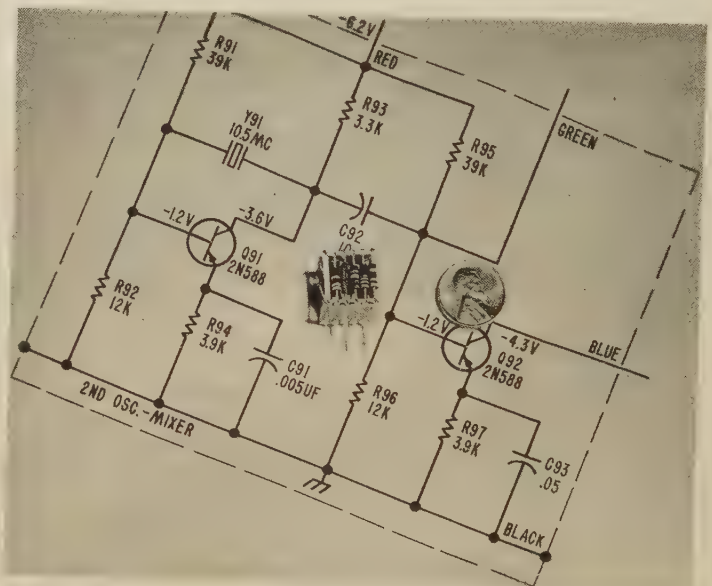


Fig. 4. Completed module shown with its schematic representation.

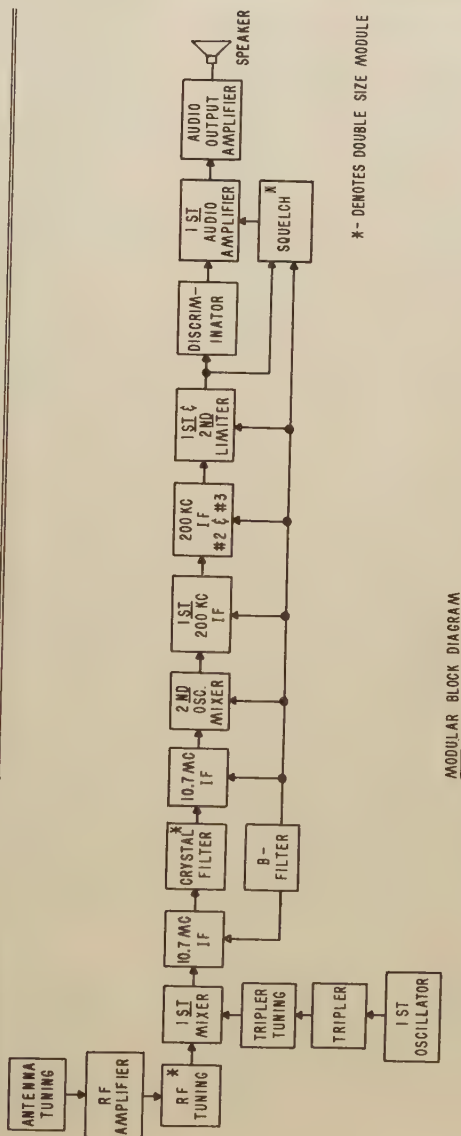
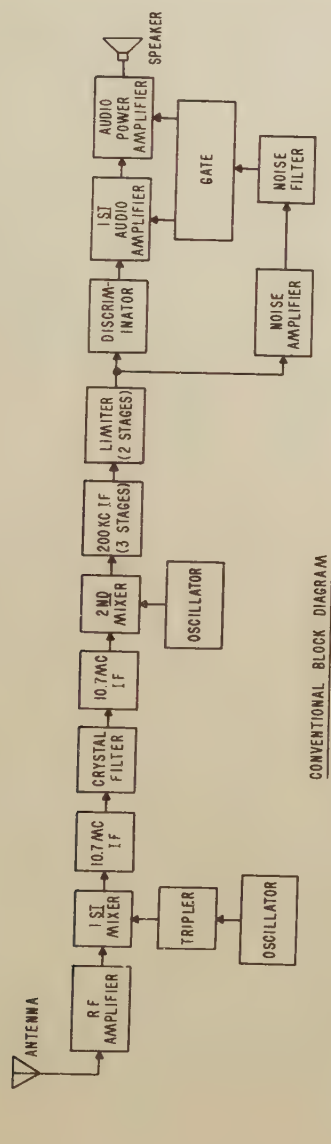


Fig. 5. Block diagram of receiver shown in conventional form and in modular form.



Fig. 6. Circuitry of companion transmitter (left) is contained in the top part of the separable case, batteries in the lower part.

PERSONAL RADIO PAGING IN THE VHF BAND

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INTRODUCTION

From the beginnings of radio, engineers have visualized the benefits of radio systems for private communications in addition to the mass broadcast radio use. There has always been a dream of radio use for private communication. Non-broadcast use of radio was one of the first uses, but many factors have limited a more general application. First and foremost, the availability of frequencies is the greatest natural limitation. However, size and complexity of radio equipment have been additional factors. The dream of the radio engineers to give everyone a radio for private communications is yet an impossibility, however, a big step in this direction is provided by personal radio paging. With one way radio paging, large numbers of users can be serviced and with selective calling the system is private. By the use of selective calling, a private page alerting call can be sent with a voice message to any one of many thousands of individuals on one narrow band VHF channel. Through the use of transistors and miniaturization, the user can now carry a small light weight receiver containing the required highly sensitive radio and selective calling equipment. With only an ON/OFF switch, a volume control, and push-to-listen button, the radio's operational complexity has been removed.

SUMMARY

This paper is a discussion of a complete VHF Radio Paging System designed to operate in the 25 to 50 mc band and 144 to 174 mc bands. The system includes small light weight receivers which are worn by the users. The receivers can be selectively called up to many thousands on the same RF channel. When the receiver is called, the user hears an alerting tone. He then presses the listen button and the voice message is received. The base station is a standard land mobile FM transmitter. This transmitter is tone modulated by the encoding equipment to send the page alerting call. It is then voice modulated to send the message. The base encoding equipment may take several forms from the simplest which can call 90 different receivers to the most complex which can call up to 7500 receivers. The base station RF equipment may take many forms and may also service two-way radios on the same channel.

THE RADIO SYSTEM

The system is designed to be compatible with all FM mobile communications system in the 25 to 50 mc band and in the 144 to 174 mc bands using either ± 15 kc or ± 5 kc deviation. The selective

calls are sent by modulation of the carrier with two tones in sequence. The tones modulate the carrier up to two thirds rated deviation. There are 88 different tones ranging from 100 to 1100 cycles spaced at 2.5%. By using all combinations that use two different tones, 88 times 87 or 7656 combinations are possible.

To insure good signal to noise in the case of the user moving through standing waves, the tones are transmitted for one second followed by a pause of carrier only for 1/4 second and the second tone is transmitted for about 3 seconds, however, it is variable up to 10 seconds. This second tone determines how long the users receiver will produce an alerting tone. A delay of 3 seconds between calls is built into the base equipment so that false codes are not produced by a combination of the last tone in the first call and the first tone in the next call. Normally, the voice message is transmitted during this time.

The system is compatible with other tone or voice operated equipment in the same system or sharing the same channel. For example, the system may also include a number of two way mobile radios with either carrier operated or tone operated (Private Line) squelch receivers but none of the transmissions to the mobile units will alert the pager receivers and the page tone or messages will not be heard in the tone squelch units which are units that remain squelched unless the correct low frequency audio tone is transmitted during the message. The paging equipment automatically disables the tone squelch signal when a page is transmitted to insure this compatibility and to allow the page tones to use full system deviation for extra signal to noise margin.

There are standard mobile radio selective calling equipments that use four tones, transmitted two at a time (such as Quik-Call), which are also compatible with this system. In these systems some of the 1400 available calls can be assigned to the pocket receivers and the same base equipment will transmit two tones in sequence with the proper timing as required on the assigned calls for the paging receivers.

The equipment will in many cases be used in an existing mobile two-way radio system when personal paging is also required. It is licensable in any two-way system, since voice messages are used. There are limited number of paging only frequencies available where the system will operate strictly as one-way paging system.

In any system the coverage prediction must be based on the effective sensitivity of the receivers as well as antenna gain and height, and transmitter power output. The encoding equipment has the flexibility of remote location from the base station antenna and transmitter sight so that the base antenna sight can be optimized. Two wire telephone line which is a standard means of remote control can be used.

THE RECEIVER

The receiver is the Motorola Handie-Talkie Pocket Receiver. (See Figure 1). It weighs 13 ounces, and is approximately 6 inches by 2 1/2 inches by 1 inch in size, complete with radio, 1 3/4 inch speaker, two resonant reeds and associated circuitry, batteries, antenna, and case. It is a standard double superheterodyne FM communications receiver with an RF section, a high IF, a selective low IF filter, a low IF limiter, a discriminator, audio and two separate crystal oscillators. The radio has 18 transistors and 8 diodes. It has a 20 DB quieting sensitivity of .5µv in the 50 mc band and 1.0µv in the 150 mc band. The paging sensitivity is 6 DB better than the 20 DB quieting sensitivity. The audio power output is .5 watts.

The radio has a built in antenna but it does not allow full coverage because of its size and proximity to the body. The antenna terminals are brought out through a jack to provide for external antenna usage. A line of accessory antennas can be used with the receiver. A helmet antenna can be attached to a metal construction hat and improve the range substantially since this makes a good equivalent to ground plane antenna with a whip acting as the 1/4 wave length above ground which is the hat. The receiver can also be attached to a mobile two-way antenna when in a vehicle with almost full mobile coverage.

The receiver has an ON/OFF switch on the front of the case and a volume control to set the voice level. It has a push-to-listen switch on the top which will switch the receiver audio to the listen condition for a voice message or to monitor any transmission on the channel. Another jack on the top along side of the antenna jack is the audio jack which is used to remote the audio output. A lapel earpiece can be used with this jack to bring the audio closer to the ear or for private usage an earpiece can be used.

Figure (2) shows the back side of the unit with access to the battery compartment. Mercury batteries can be used as shown or a rechargeable nickel cadmium battery can be directly interchanged. The battery contacts on the rechargeable batteries are brought out on the bottom of the unit so it can be recharged in the radio.

Figure (3) shows a series of battery chargers for recharging batteries. The unit on the left will take 10 radios and charge the batteries in the radios. It can be expanded to a total of 20

radios. The unit in the center will recharge a single receiver. The unit on the right will handle up to 23 batteries separately for uses where the radio must remain in service. In all cases, the charging action is the same. That is, a red light lights when the battery is placed on charge. The charge cycle takes approximately 16 hours, but the battery can be left on charge indefinitely without damage. The chargers are designed to deliver a constant current charge and limits the charge or over charge, as the case may be, to a safe value.

In Figure (4) a view of the upper back section of the radio with the top opened for servicing shows the basic radio and selective calling circuits to be a separate section of the housing from the battery compartment for protection of the radio. Here twenty separate printed board modules are mounted in a manner that exposes all the basic radio connections for service. The nineteen modules of the basic radio are housed in a cell crate with the 455 kc filter. The crate is built around the speaker magnet. Above the radio, two resonant reeds can be seen along with the printed board module holding the selective calling circuitry. Figure (5) shows a blow up of the receiver with a few of the basic parts. The speaker and cell crate can be seen. This cellular housing for the receiver modules is contoured around the speaker magnet and is constructed of plastic. This is then copper and gold plated to form a shield and solderable ground. The two resonant reeds are shown on the left with the two jacks. On the right the rechargeable nickel cadmium battery can be seen. A few of the radio modules are shown below the cell crate. The component density in this radio is as high as 250,000 parts per cubic foot in some of the amplifier modules and is over 125,000 in the total radio section using conventional components. Figure (6) shows a service man placing a module into the cellular housing. A section of the cellular housing is cut away to show another module in the housing. The cellular housing is mounted in the radio case on sponge rubber pads to protect it from shock and vibration and thus provides a very rugged unit.

The block diagram of the receiver is shown in Figure (7) for the 150 mc receiver. The 50 mc receiver is quite similar with an RF amplifier stage added and a 5 mc high IF in place of the 12 mc IF. From this diagram the signal path can be traced. The RF carrier passes through the RF selectivity to the first mixer where it mixes with the third harmonic of the first crystal oscillator which is multiplied in a diode. The resulting 12 mc IF frequency is amplified in the two high IF amplifiers and then it is fed into the second mixer which is energized by the second crystal oscillator. The result is 455 kc which is amplified in the second IF after passing through the 455 kc filter. This filter is the main radio selectivity, and can be changed to shift from ±15 kc operation to ±5 kc operation and provides for the rejection of the adjacent channel which may be 20 kc or 40 kc away in the 50 mc band or 30 kc or

60 kc away in the 150 mc band. The audio modulation is detected in the discriminator and used to drive the two resonant reeds through a three stage amplifier. When the right tones are received in the right sequence, the reeds pass a charge through their contacts from the first reed to the second and then to the bias circuit of the audio driver stage. This stage is prepared to oscillate since the necessary feedback is connected through the push-to-listen switch. When the bias is developed, the audio driver stage oscillates. This is amplified by the push pull audio output stage to about .7 of a watt. Some of the energy from the alerting tone output is rectified and fed back through the second reed contacts as bias so that the oscillations will continue as long as the second reed is vibrating. When the push-to-listen switch is depressed, the audio from the discriminator is connected to the audio driver stage, which has its feedback removed and its bias restored so that the audio can be heard in the speaker.

THE BASE STATION EQUIPMENT

Figure (8) outlines a typical small system for radio paging. Note up to 360 receivers may be called in this system. The control equipment consists of a console and microphone as shown in Figure (8). The control console contains all the necessary equipment to encode the transmitter and can be remoted from it as shown. The VHF Base Station shown is a high powered two-way mobile transmitter housed with an associated receiver and power supply. The variation this base station transmitter and receiver may take are not restricted to the paging system. Many types of equipments and features are possible. This transmitter and receiver may also be controlled by additional remote or local console for regular push-to-talk two-way use with mobile units. One position must be designated as the control point and it must have supervisory control of the system, however, the type of modulation from these other remote dispatch points may vary. They may be transmitting and listening to mobiles with a tone operated squelch, or to units with selective calling receivers, or to units with standard noise operated squelch.

Figure (9) shows a radio paging system with a large capacity encoding system. In this system, the encoding equipment is housed in a separate encoder cabinet. This encoding equipment can be controlled by up to eight dispatch points. At these points the operator has a "fast dialing push button box" as a control unit and some type of microphone or a telephone handset. Each box also has a speaker or earpiece and volume control to monitor the channel when the listen button is depressed. This equipment enables high speed selection of the paging code. The output of the encoder cabinet can be connected to the base directly or by two wire lines. Again the basic transmitter and receiver can take many forms and also can have additional remote inputs other than from the paging control units.

SMALL SYSTEMS

For small systems a 90 call control console is used which is shown in Figure (8). This provides the necessary tone outputs to drive any one of a large array of basic transmitters which have from 25 watts up to 1/4 kw of RF power output in either low or high band (50 or 150 mc). It has two rows of 10 buttons which activate two of ten oscillators in sequence to provide the modulation. After the buttons are depressed, the transmitter is automatically turned on and the tones are transmitted automatically with the proper sequence and timing. After the last tone, the indicator light goes off and the operator can depress the mic switch to send a voice message. This console has provisions for monitoring the channel with a speaker, a volume control, tone coded squelch disable switch, and listen switch for standard squelch systems. A transmit switch and reset button are also included. The reset button can be used to reset in case of an error in selection of the code. Up to four consoles can be paralleled to provide up to 360 calls. This console is similar to the console used for regular two way mobile selective calling which can be used compatibly with the pocket receiver paging system.

Figure (10) outlines a block diagram of the console showing the various functions of this equipment. The ten oscillators in the console are stabilized in frequency by resonant reeds in the feedback circuit. This maintains frequency within $\pm 1.2\%$ over the temperature range and with time. The output of each oscillator is fed into the timing circuit through the twenty push buttons. The first tone is transmitted by the timing circuit for one second and is selected on the top row. The second tone is transmitted for 3 seconds and is selected on the bottom row. The buttons latch down showing the number selected for the duration of the first and second tone transmission times. The timing circuit keys the transmitter as well as times the tones. The transmitter can also be keyed by the push-to-talk switch on the microphone. The audio for the microphone and the tones from the timing circuit are fed into the line amplifier when used in remote control operation. They are amplified and fed into a 600 ohm two wire line to the remote transmitter sight. The transmitter keying is also fed down this line as a DC signal. The squelch disable or tone operated squelch disable is also fed down this line as DC. The audio from the base receiver, which is used in systems where paging is operating on the same channel as two-way mobiles, is returned on this line to the speaker and volume controls. This enables the dispatcher to monitor the two-way call as well as the facility to monitor for a clear channel before a paging transmission.

LARGE SYSTEMS

For larger systems the encoding equipment can handle up to 7500 different receivers on one RF channel. The basic equipment is housed as is shown on Figure (9) and (11). Two of the remote

control units are shown in Figures (12) and (13). One has a handset for a microphone and speaker, and the other has a flexible arm microphone with speaker built into the control unit. Note these control units contain numbers 1 to 9 and 0 plus a button marked "L" to monitor the channel, and another marked "T" to key the transmitter. The "T" button is used to send a voice message or for use in normal two-way radio communications with a mobile unit on the channel. A red "transmitter on" indicator light is mounted on the unit and a green light is used to indicate that the paging tones have been transmitted and the channel is clear to send a voice message.

Figure (14) is a block diagram of the high capacity system. On the left hand side two encoding equipments are illustrated. The upper unit is the control position and has supervisor control over the base station so that it can over ride the other position as required by the FCC.

The output from this box to the encoding equipment cabinet is connected through a twelve conductor cable. As the first button is pushed on the control box, a code is started in the system. The line input relay transfers the code selection to the next relay and as a primary function locks out all other line inputs by removing this supply voltage. An indication of busy goes back to the other control units in the form of a red light. At this point all other units except the control point unit are locked out until after the control unit in use is reset. There may be up to eight other inputs in the standard equipment.

The code selection information next goes to the first transfer relay which feeds the binary storage bank. The transfer relay transfers the code selection after each digit to the next transfer relay and storage bank. This insures that only one selection is made on each binary storage bank. The binary storage banks each consist of four fast acting relays which will energize in accordance with the code selected. Using a binary number system in these banks allows the use of four relays to indicate one of ten digits. More digits are possible, with four relays, however, only combinations with one or two relays pulled up are used. The binary approach allows the use of not only four rather than ten relays in the bank for each digit, but also it allows the use of only four wires from the control boxes for each of ten digits selected on the push of each button. The primary reasons for using the storage system is to allow high speed selection of codes. In this system, the code equipment can be called at a rate of 600 digits per minute. The push buttons are mechanically interlocked so that only one digit can be pressed at one time. This system will not jam if more than four buttons are pressed, the first four will be used to transmit a call. If less than four are pressed the system will reset itself after one minute. The transmitter is automatically keyed for 15 seconds after the tones go out so that the voice message can follow without further button pushing. The last protection to

complete the human engineering at the push-button control unit is a reset button to reset the system if the operator makes a mistake before the fourth digit or to stop any of the timing cycles.

As the second button is pressed in the code selection, the code is started in the second binary storage since the first transfer relay has transferred the code signal path to this storage. This process is repeated in quick succession until the fourth and last number is pushed on the control box. Four binary storage banks are used in this system with one for each of the four digits in a call and then two banks are used to set up one tone of the two tone transmitter modulation. The four binary storage banks are swept by six stepping relays to select the proper oscillator tone as determined by the code in the storage bank. The steppers are used to accommodate the large number of selections and contacts needed in converting from the binary system to the decimal system for tone selection. They cannot be driven directly from the push buttons because they are too slow for the fast input that is desired. They would also require more wires in the control cable since they would have to operate on a decimal system. The first stepper starts at the end of the first binary storage input and it is likely to be set long before the last button is pushed, thus the tones can start being transmitted immediately. One stepper is used with the first binary storage bank and selects one row of 10 oscillators in a 10 by 10 matrix. The second binary storage bank is serviced by two steppers to select one row of 10 rows in the other leg of the matrix. The rows 1 to 5 are serviced by the first stepper in the second binary bank and rows 6 to 10 are serviced by the second stepper. Each stepper servicing the second bank thus has five rows of contacts with 10 contacts per row. The steppers in both cases step at their self-interrupted rate seeking the open contact of the binary code. The second and third banks of steppers are serviced in like manner by three stepping switches to convert the binary code to two of 88 tones.

The two tones selected are fed to the sequence timer. Two control tones are sent as a pair as the first operation of this timer. These perform the function of transferring the transmitter input from the normal voice input to the tone input and they also disable the tone at the transmitter that is normally transmitted in tone controlled squelch systems. The sequence timer transmits these control tones for .5 seconds and then transmits the paging codes in their proper sequence. The timer as shown is energized by the timer relay which is activated by the last stepper. The tones are transmitted along with DC to key the transmitter by the line amplifier chassis. The voice delay timer holds the transmitter on for a pre-selected time so that the operator can send the voice message without depressing additional buttons. The transmitter can also be manually keyed at the push button box. The transmitter control timer is the timer that resets the system if the code input is not complete or for some other

reason the system should be locked up. This resets the system and turns the transmitter off. The voice delay and other action can be removed by the reset button and manual reset relay.

At the transmitter sight a remote control chassis is used to provide operation over two wire line. The received messages from the base station returns to the control point via the two wire line and through a change over relay in the line amplifier.

CONCLUSION

This paging system is designed to further utilize the land mobile frequencies by providing one-way paging. The equipment is compatible with the vigorous standards required in these crowded channels. It is also flexible in that it can operate in a system with mobile equipment and has control equipment for small systems as well as the largest systems of up to 7500 calls. It is designed for maximum user efficiency by providing both the paging alert call and a voice message to the user.



Fig. 1. Handie talkie pocket receiver.

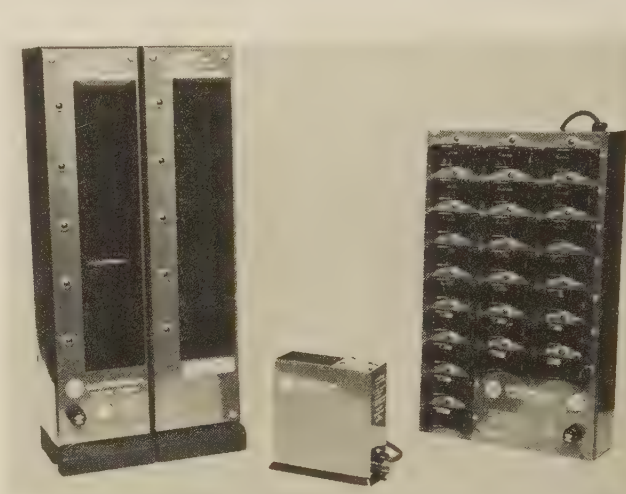


Fig. 3. Three types of battery charges that are used to recharge nickel cadmium batteries.



Fig. 2. Receiver battery compartment.



Fig. 4. Pocket receiver with radio section opened for servicing.

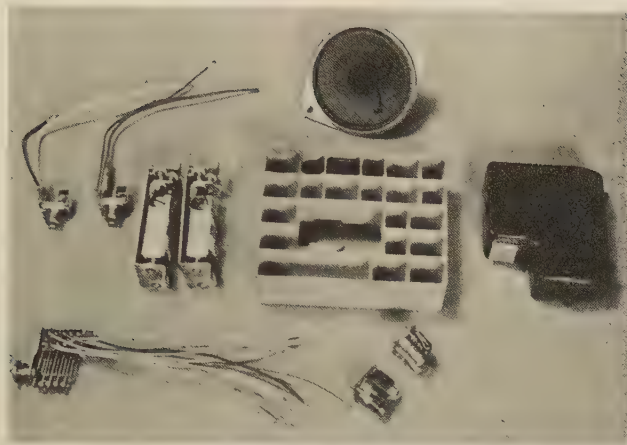


Fig. 5. Basic parts of the pocket receiver.



Fig. 6. Service man placing module in the pocket receiver.

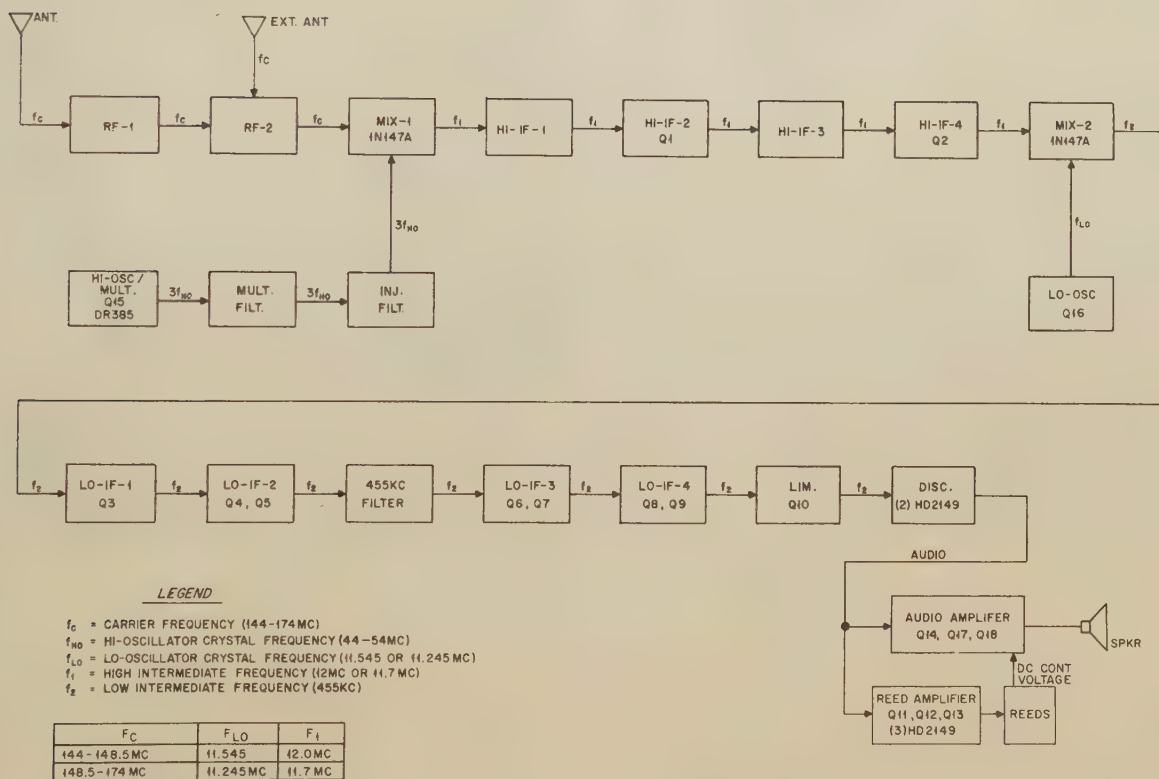


Fig. 7. Pocket receiver block diagram.

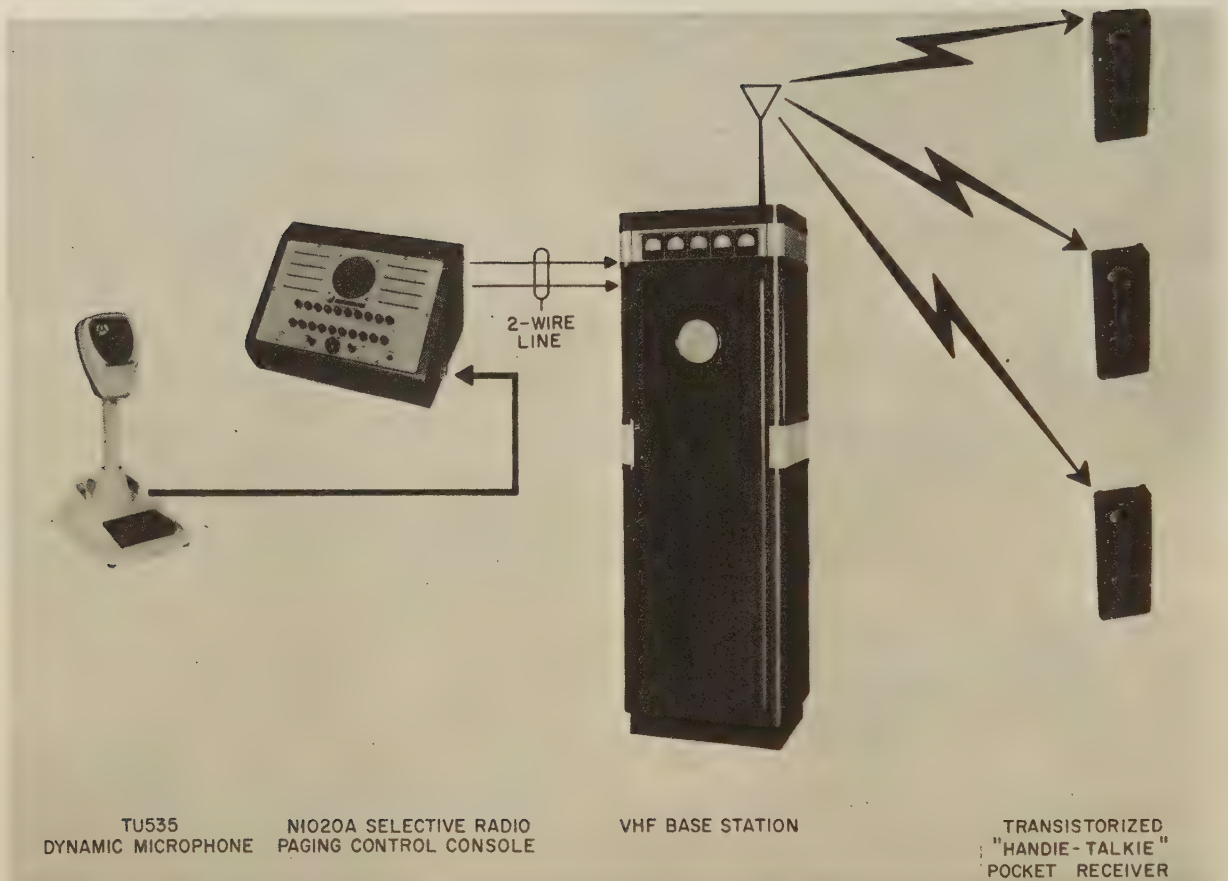


Fig. 8. Small system selective radio paging system.

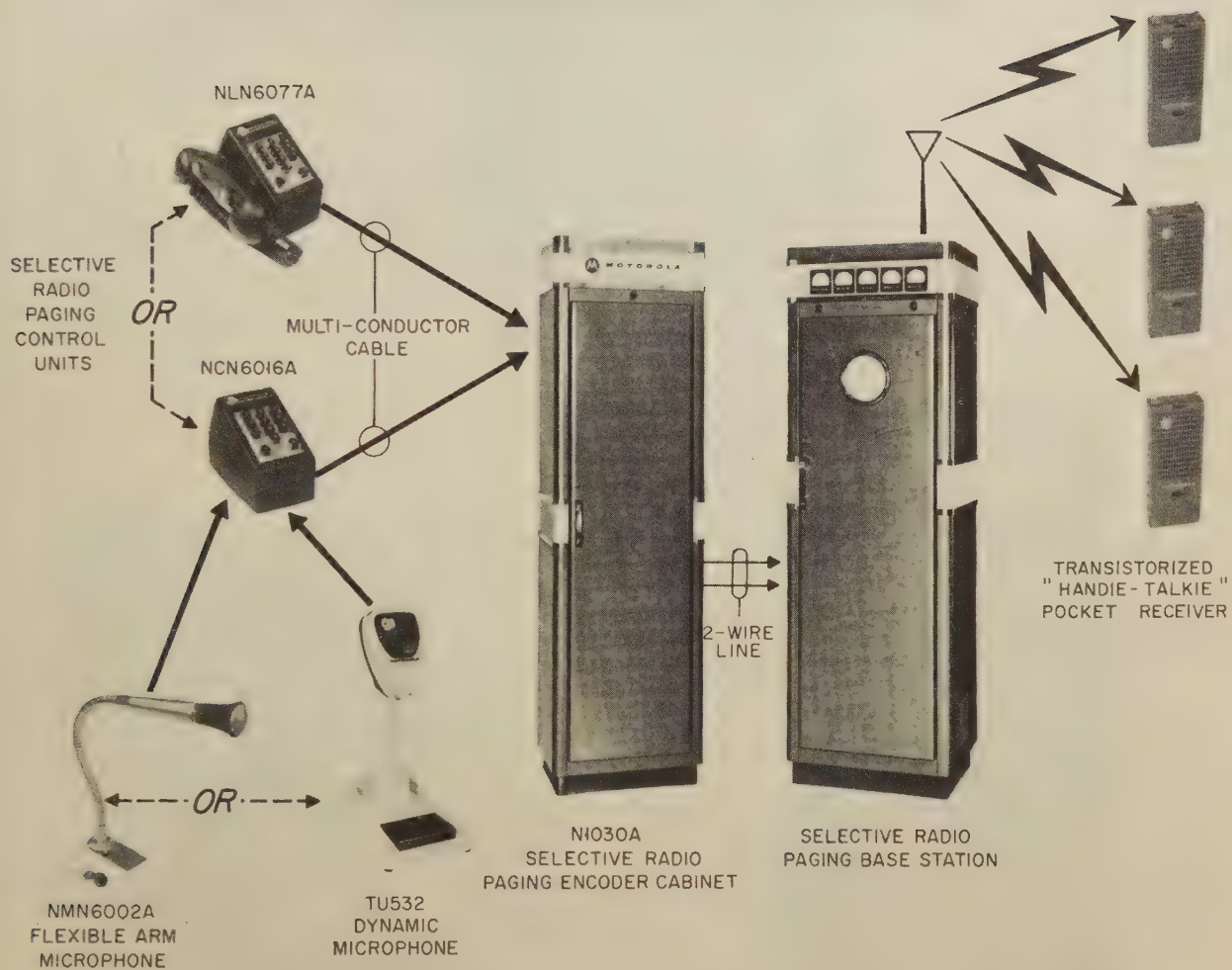


Fig. 9. Large system selective radio paging system.

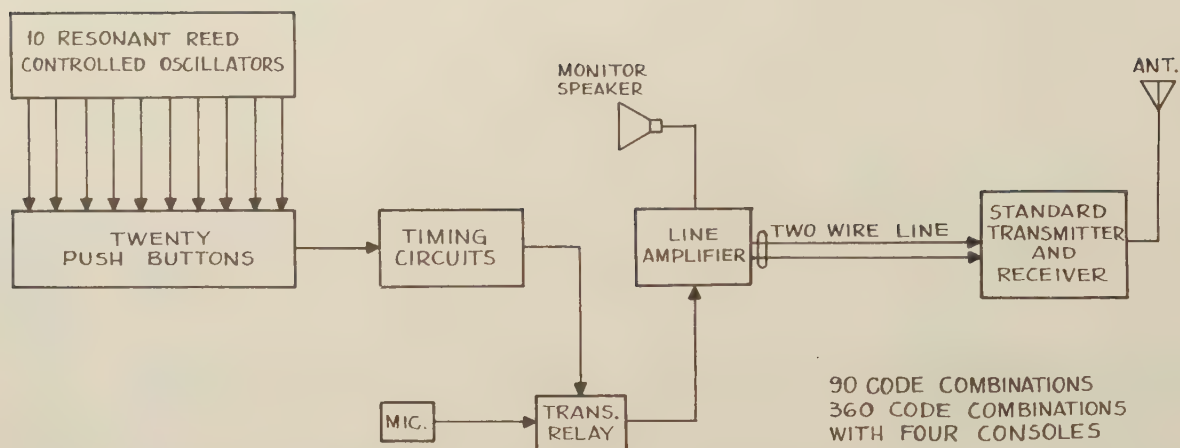


Fig. 10. Block diagram of small system paging control console.

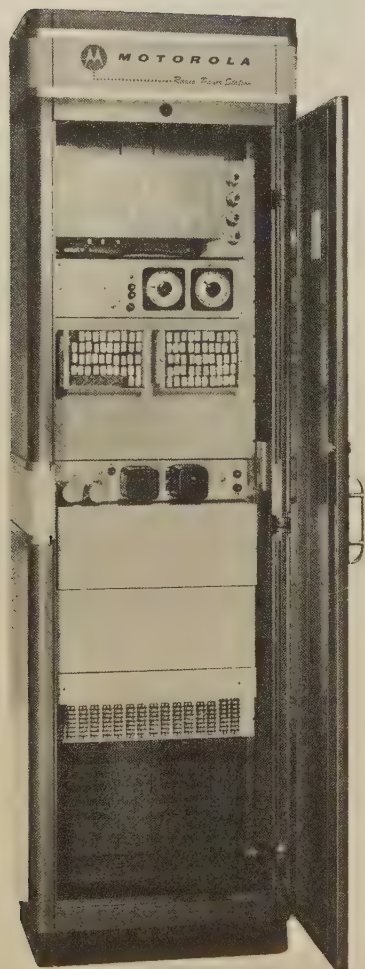


Fig. 11. Inside view of the large system selective radio paging encoder cabinet.



Fig. 12. High capacity system control unit with flexible arm microphone.



Fig. 13. High capacity system control unit with hand set.

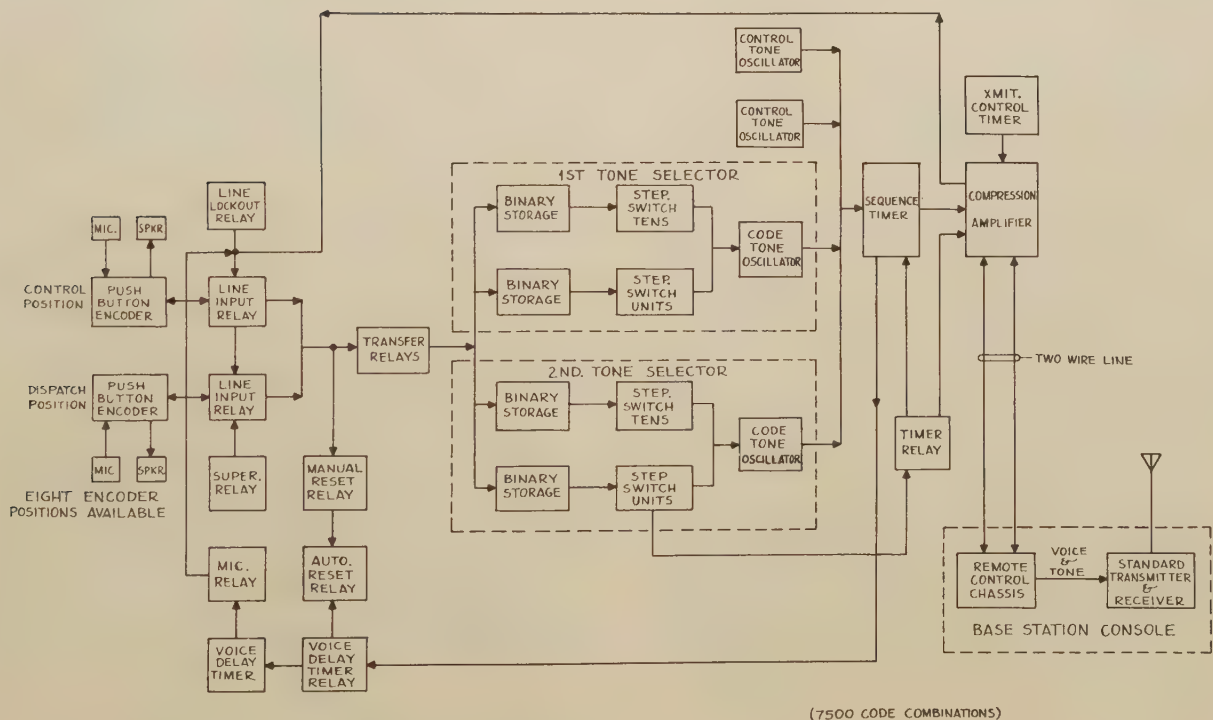


Fig. 14. Block diagram of high capacity encoding equipment.

POLICE AND FIRE DEPARTMENT COMMUNICATIONS CENTERS
A SYSTEMS APPROACH TO THE CONTROL CONSOLE AND RELATED FACILITIES

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Summary

Police departments, and similar organizations which require communication with mobile units are confronted with similar problems. The equipment provision for a typical installation is discussed with particular emphasis on the facilities which are provided in the communications console.

A description is given of the modular units such as amplifiers, control units, display units and signal actuated recorder which have been developed for this application.

Introduction

Early in 1959, Westrex Corporation commenced a survey into the requirements for equipment to be used by police and fire departments in conjunction with their radio-communication systems. It was recognized at an early date that contributions could be made towards increasing the efficiency of operation of such systems by designing a completely new line of equipment specifically for this type of application. It was decided not to limit the development of new items to those required purely for the control and monitoring of the radio channels but to include auxiliary equipment such as signal-actuated, multi-channel recording systems, surveillance equipment, status control devices and visual display units.

The development program which was undertaken was directed initially at providing equipment which would meet the general requirements of both police and fire departments, and this portion of the project is essentially complete. There are several problems of control, however, which are peculiar to certain installations such as fire departments, and work is now proceeding on recommendations for a modernization plan for these and other specific applications.

The primary design consideration was the requirement that police radio equipment must be on the air at all times and therefore that no lost transmission time can be tolerated for repair or replacement. As a result, modular plug-in units have been developed and incorporated into systems wherever it has been economically practical to do so. In other instances such as in the case of power supplies, spare equipment and automatic transfer facilities have been provided.

Versatility and adaptability were other important considerations, because it was obvious that no two installations, large or small, were likely to require precisely the same equipment and it will be noted in the following discussion that the same plug-in electronic modules are used in several different applications. This is also true in the case of specially-developed equipment such as the signal-actuated recorders, discussed below, which can provide facilities ranging from simple, continuous, single-channel recording units to complex, multi-channel, voice operated units employing automatic, fail-safe, transfer features.

Requirements of Police Department
Command Center Installations

Before attempting to describe the equipment required for a typical installation, it would be well to review briefly the general operating procedure which is used. The amount of equipment supplied and the number of persons employed will naturally vary widely from city to city; however, the principles remain very much the same.

When a call is originated by a member of the public, it is received at a complaint-desk position where the details are entered on a card.

In the event that it is necessary to dispatch a police vehicle, the card is routed to a dispatch console where the operator has control of one or more radio channels. In the smaller organizations, this is usually accomplished merely by the handing of the card from one person to another, but in the larger installations, it is customary to provide a conveyor-belt system between the two types of position.

It is the responsibility of the operator at this dispatch console to establish radio contact with the police unit to relay details of the complaint, and to maintain a record of the status of the various vehicles under his control.

In certain installations, it is customary to record all telephone and radio communication conversations on some type of tape or disc recorder.

This is basically all that is involved in the operating procedure. There are, of course, other activities not covered in this brief review, some of which will be brought to light in the subsequent description of a large police department installation.

Description of a Typical Large Police Department Installation

A typical example of a large command center is that designed to meet the requirements of Dade County, Florida; Figure 1 is a floor plan of part of the area devoted to the communications section of the police department.

In the large room it will be noted that there are twenty complaint desks, and between the two rows of ten such positions passes an IBM-card conveyor belt system. As complaints are received, the details are entered on IBM-type cards; the cards are then stamped with the time and date and inserted into one of the seven tracks of the conveyor belt system. Each of the seven tracks corresponds to one of the seven dispatch consoles, and each console operator has a specific area of the county allocated to him.

To the rear of the complaint desk position is a map display board which provides space for mounting maps of five different areas of the county. The facilities provided by the map display system are discussed in a separate section below.

Dispatch Consoles.

The dispatch consoles are located in seven rooms separated from each other and the complaint desk area by plate glass partitions.

Figure 2 is a front view of one such dispatch console, and each of the seven is essentially similar apart from the fact that six of the units are arranged in pairs back-to-back and that alternate units are virtually mirror images one of another. Moreover, one console is considered the primary position and has both additional equipment and additional facilities. Figure 2 is marked to show this additional equipment.

The overall configuration of the shell and layout of controls is such as to provide maximum accessibility and ease of operation. All controls which require frequent actuation are within immediate reach of the operator.

Transmission-Receiving Controls. At the center of the desk section is an inclined control panel upon which are mounted ten transmitter-receiver control keys, five on either side of a small area occupied by compression and volume indicator meters and an alert tone key. Seven of the transmitter control keys are connected to the radio transmit-receive circuits currently available, and an eighth provides for simultaneous operation of all channels under emergency conditions. The remaining two keys will provide for future expansion of the system. Any channel can be controlled from any of the seven consoles.

In keeping with the design policy, the transmitter-receiver control keys are plug-in modular types, each controlling a particular transmitter and receiver channel. As shown in Figure 3, each unit consists of a five-position switch and lamp indicator unit. In position 1 of the switch, the associated monitor speaker is disconnected on that console position but remains connected to the receive channel on the remaining positions. Circuitry is provided also to sound an alarm at each console in the event that the control keys pertaining to a particular channel are all actuated to position 1. This is to protect against the possibility of all operators disabling the monitor circuit of a specific channel. Position 2 of the switch connects the monitor speaker to the receive line. Position 3 is employed in conjunction with a sequential two-position foot-switch. In position 4 of the control switch, a d-c signal is applied to the control pair to operate the squelch-disable relay to permit monitoring the channel before transmission. Position 5 applies a d-c signal to the control pair to actuate the transmit relay circuitry in the transmitter. In all but positions 5 and 6, the transmit control key remains in the position to which it is operated.

The footswitch in effect duplicates the last two positions of the key switch; that is, it first actuates the squelch-disable relay circuit to provide for monitoring and then permits transmission on the line. Any one transmission line, or any combination of them can be operated simultaneously by the footswitch if the corresponding channel control switches are in position 3.

The lamp section of the transmitter control switch module contains six sections which are engraved and illuminated from the rear in different colors for different conditions. A section engraved "Transmit" is illuminated in green when any console operator actuates his transmitter control key to position 5 or when he uses footswitch transmission. Circuitry which will be described later prevents a particular channel from being seized by more than one operator at any one time. An audible warning is given also in the event that a second operator attempts to transmit on a channel which is already in use.

When a transmit control key is operated to position 1, a section of the lamp assembly which is engraved "Off" is illuminated in red at that console position only. The audible alarm circuit discussed previously operates in the event that the transmitter control key associated with a particular channel is operated to the "Off" position on all seven consoles.

The prime dispatch console position has been mentioned previously, and one of the additional facilities incorporated permits the operator to take overall control of any or all channels under emergency conditions. As an indication to other operators, the lamp assembly on each channel control switch incorporates a section engraved "Override" which is illuminated when this condition exists.

A visual indication is given also when the transmitter control is in the position for footswitch operation by illuminating in amber another section of the lamp assembly which is engraved "Ready."

The two remaining sections of the lamp assembly are devoted to carrier indications for the two receivers associated with a particular channel and are engraved respectively "Rec A" and "Rec B"; these are illuminated in white.

In the transmitter control key panel are located the volume indicator and compression meters to indicate respectively the level and degree of compression of the audio signal which is being transmitted to line. The transmission circuitry and compression amplifier with which they are associated are described in subsequent paragraphs.

To draw attention to the fact that a particular transmission is of greater than usual importance, it is common practice to precede the transmission with a short burst of tone which is commonly referred to as an "alert" tone. A transistorized single-stage oscillator in the form of a plug-in module has been designed for this purpose. The alert-tone operating control is located beneath the meters on the transmitter key control panel.

In each console, one monitor speaker is provided for each channel and these are arranged horizontally across the top of the turret section. Each is provided with a volume control with a minimum loss position to prevent the monitor circuit from being turned completely off.

Status-Indicating Equipment. In a number of the Command Center installations visited in connection with the investigation into the facilities required, it was noted that the record of status was maintained for vehicles and personnel purely on the basis of written information. While this is usually fairly satisfactory where only small numbers of units are involved, it becomes progressively more difficult to locate a particular vehicle as the numbers of units and transmissions increase. It is further desirable in some of the larger installations for personnel in different areas to have an overall picture of the situation at any particular moment.

It was decided therefore to design a switch which could be employed to provide a visual indication of various conditions of status. It was further agreed that this type of switch should be in four versions; the first would be purely a mechanical device with no electrical connections; the second would be a similar unit but incorporate contacts for connection to a remote display unit; and the third and fourth versions would be similar to the above two units but be provided with an extension shaft to permit operation from either side of a pair of consoles which were arranged back-to-back. Discussions with certain police departments indicated that they favored plug-in type modules for this purpose.

Referring again to Figure 2, it will be noted that in the vertical turret section, there are two banks of status switches arranged one on either side of the central panel. In this particular installation, the consoles are of the double or back-to-back type, and the switches can be operated from either console. Such a switch is shown in detail in Figure 3. A section of the knob is cut away so that different colored plastic segments on the front mounting plate are displayed as the knob is rotated, to correspond to different status conditions.

Electrical connection is made from the switches to visual display modules which are mounted in cabinets both in the dispatch console area and in the complaint-desk room. A sketch of the lower part of one of these cabinets is shown in Figure 4 with one of the light-display box units removed to reveal one of the six status lamps which are provided in each module. Each cabinet in this installation accommodates 50 such units.

Due to the fact that discoloration of bulbs of the painted type has occurred over a period of time in similar use, it was elected to employ standard clear bulbs and to provide a filter of a different color in front of each bulb for this application. On the front of each module is a translucent plastic screen which is illuminated from the rear by each bulb and upon which is silk-screened an identifying code number. This screen of each module is interchangeable and the code numbers employed correspond to similar numbers on the associated status switches.

In the Dade County installation, the rotary switches and the corresponding indicators in the status cabinets are identified with various supervisory officers, members of the detective force and other personnel not associated with patrol duties. The colors are coded to indicate the whereabouts or the availability of these personnel. A status-display cabinet for each console is mounted near the ceiling in the dispatch room on the wall separating the dispatch and complaint-desk rooms. Either operator of a double-console can see his own cabinet and

that of his partner through the glass partition. These status-display cabinets are duplicated in the complaint-desk room to provide personnel there with this same formation.

Next to the banks of 50 status-control switches on the vertical back panel of the dispatch console shown in Figure 2, can be seen the IBM-type card units. Each of these consists of four tiers of twenty-five numbered slots designed to accept the IBM-type complaint cards received via the conveyor belt from the complaint area. After the card is stamped in the automatic time and date unit mounted on the console, the dispatcher reads the complaint, dispatches a vehicle, and then inserts the card in an appropriate slot. Each of these slots corresponds to a patrol car, a motorcycle, or a similar unit and is coded with an identifying symbol. A card inserted in a slot on one side of a double console will protrude from, and can therefore be withdrawn from, the opposite side. Thus, as in the case of the rotary status switches, a single operator can operate the facilities of both consoles. The insertion of individual cards in particular slots operates a single-pole, two-position switch connected to lamps in 50-unit, status-indicating cabinets similar to those connected to the rotary switches. These are mounted on the wall in the complaint-desk room, but instead of six lamps in each indicator module, they contain only two: one to indicate that the vehicle is available for use and the other to indicate that it is not. This same information appears in the illumination of either of a pair of red and white lights on the large county map.

Miscellaneous Dispatch Console Provisions. A number of other provisions are shown in the sketch of Figure 2. A twelve-station inter-communications unit, mounted in the vertical wing section of the console, provides selective communication with the other six consoles and with five of the complaint-desk positions. The vertical center panel at the back of each console contains a cyclometer-type clock and a set of six inter-console emergency pushbuttons to signal the other consoles, by visual and audible indication, that a specific channel should be released for an urgent transmission.

A headset monitor circuit is available at a jack on each console to permit monitoring of any one of the radio channels by operation of a ten-position selector switch.

Prime Position Console. The prime console includes several control facilities not found on the other six consoles. Among these are a Master Alarm Detector to sense the presence of alarm tones indicating failures in the radio equipment and to identify and locate these failures, and a Motorola "Quik-Call" unit to provide radio-operated, remote-control, selective signalling facilities. In addition, the Master Position Console, as mentioned earlier, is provided with

override facilities which enable the operator to assume emergency overriding control of all radio transmission channels.

Map Display and Storage Unit.

Figure 5 is a sketch of the map display and storage unit located in the complaint-desk room of the Dade County Police Department. It will be seen that although five map sections are provided, only three are to be used initially, the other two being provided for future expansion. Each map is mounted on a sheet of aluminum approximately 46 inches wide by 76 inches high, and perforated to admit 100 pairs of red and white lamps, each pair representing a dispatch vehicle. One or the other of these lamps is illuminated at all times to indicate status in that area. When a dispatcher sends a vehicle on a specific mission, he inserts an IBM incident card in a slot on his console thus operating switches which extinguish the white lamp and illuminate the red one.

As shown on the center map in Figure 5, six clear plastic overlays mounted on rollers can be pulled down over the map areas. These overlays are marked with the outlines of different "beat" areas, any one of which can be selected.

The three-section base of the unit provides storage space for additional maps.

Transmission Circuit. Figure 6 shows in simplified block schematic form the transmission equipment which is supplied with the dispatch console portion of the system and is typical of the approach employed in other similar installations.

The output of the dynamic-type microphone at each console is connected to the input of a transistorized preamplifier and compressor amplifier. Both amplifiers are of the plug-in, etched circuit board type construction, and the compressor amplifier can be used for other applications independently of the microphone preamplifier section.

The preamplifier unit has been designed to operate from a 30-ohm microphone source and with an input of -70 dbm provides an output level of -20 dbm into a 10,000-ohm load impedance. Transformer input is provided and two common emitter stages with negative feedback are used to provide gain stabilization. Standard low-frequency dialogue equalization is incorporated.

The compressor amplifier unit may be operated directly from the output of the preamplifier section or alternatively may be bridged across a 600-ohm line. The input impedance is approximately 30,000 ohms and the rated output +33 dbm. The overall gain of the two units from microphone input is 103 db and a fixed compression characteristic of 30:3 is employed. A threshold

adjustment is provided. Attack and release times are approximately 2ms and 3 seconds respectively.

The compression circuit is at the input to the amplifier, since compression must be accomplished at low voltage level in order to minimize distortion due to the necessarily non-linear diode characteristics. Following the compression circuit are three voltage gain stages, and a driver stage which is transformer coupled to the push-pull class B emitter follower-output stage. The amplifier is designed to operate from a 28-volt supply.

Compression is accomplished by rectifying a portion of the output signal and applying a d-c bias voltage to two diodes which are connected in series in the base circuit of the amplifier input stage as indicated in Figure 7; the input audio signal is applied through a 25 K ohm resistor to the junction of the two diodes. This point is also coupled to the input stage base through a 1 μ f condenser C2. The 25 K ohm resistor and diodes CR1 and CR2 in parallel act as an audio voltage-divider and by application of the correct d-c bias potential, the diodes are confined to the logarithmic portion of their dynamic resistance characteristic, thus providing an extremely flat compression slope.

Diode CR1 acts as a buffer between CR2 and the d-c control voltage. Since the d-c path is through the two diodes in series, the direct current flowing through them is the same. The resistance varies simultaneously and the loading effect on CR2 by CR1 is equal to the instantaneous resistance of CR2. In practice, the diode resistances vary from approximately 30 K ohms to 5 K ohms. If CR1 were eliminated, it would necessitate injecting the d-c control through a resistance of at least 300 K to prevent loading effects on CR2, and would greatly decrease the efficiency of the d-c control voltage supply.

CR1 also serves another very important function. It will be noted that the diodes are connected in parallel with reference to the a-c signal component, but have opposing polarities.

The instantaneous a-c signal voltage which appears at the anode of CR2 is applied simultaneously to the cathode of CR1; thus the net a-c swing on the non-linear diode curve is zero, thereby providing reduction of distortion effects.

Referring again to Figure 6, the output of the compression amplifier is connected to the console volume-indicator meter and also to dividing network, the outputs of which are, in turn, connected through the contacts of the transmit-receive control relays and through level adjusting pads and line transformers to the transmitter lines.

Transmit-receive relays are provided on the basis of one per operator per channel, and circuitry is arranged so that as soon as one operator has

operated his transmitter control key on a particular channel to the transmit position, the appropriate transmit-receive relay operates and prevents any other operator from gaining access to that channel. Operation of transmit-receive relay in turn actuates a relay to apply a suitable d-c potential to the line to key the transmitter.

Until such time as a transmit-receive relay is operated, the associated output of the dividing network is terminated in a closing impedance.

Other functions performed by the contact of this relay are illumination of the transmit indicator lamp in each associated control key module and preparation of a circuit for actuation of an audible alarm in the event that a second operator attempts to seize the same channel.

The outputs of the secode and alert tone oscillator circuits are connected to the input of the compression amplifier unit. The alert tone oscillator is a fully-transistorized, modular unit of the single-stage, phase-shift type and provides a 1000-cps tone. The output level is nominally zero dbm with distortion of less than one percent. The unit includes a continuously variable gain control.

Referring to Figure 6, the circuits from the receiver are connected through the line transformer to compressor-amplifier units identical to those described for the transmit circuit, and the outputs are routed through power amplifiers, transmit-receive control relays, and level controls to the various monitoring speakers. The contacts of the transmit-receive relays in this instance disable the local console monitor speaker when transmission is taking place from that console.

Control Circuit and Miscellaneous Provisions.

Figure 8 is a simplified schematic to indicate the approach to the design of the relay switching circuitry and to show the miscellaneous provisions which are incorporated. All relays employed in the system are of the plug-in type.

For clarification, each d-c function is shown in a separate block, and although, for example, the channel control switch has five positions, the drawing is simplified further by indicating only those contacts of the switch applying to the particular function under discussion.

Monitor Off Position. The block labelled in this manner shows the circuit functions which result in the disconnection of the associated monitor speaker and completion of a circuit for illumination of the indicator lamp. Also a circuit is prepared for the operation of the alarm buzzer in the event that all other operators also actuate their associated channel switches to the "monitor off" position. For simplification, only one additional operator circuit is

indicated.

Although not shown on this drawing, the second position of this switch results in the restoration of the monitoring circuit and extinguishing of the lamp.

Ready Position. This block indicates the circuits made directly through the contacts of the switch and indirectly through relay contacts to prepare the channel for transmission by subsequent operation of the foot switch. It also completes a circuit for illumination of the "ready" indicator lamp.

Check Position (Squelch-Disable). In this position of the switch, suitable voltages are applied to the control line to disable the receiver "squelch" circuit and permit monitoring that particular receiver prior to transmission.

Transmit Position. Actuation of the switch to this position results in the operation of the appropriate Transmit (TRC) relay. One contact completes the circuit for the transmit indicator lamp. Another contact prepares a circuit for operation of an alarm buzzer in the event that the second operator simultaneously operates his channel key to the same position. Contacts of the TRC relay will operate a second relay to apply the required signalling voltages to the telephone line through a filter circuit and thus actuate the transmitter. Another pair of contacts removes the termination from the dividing network and connects the appropriate channel to the network output.

Protection is afforded against the second operator transmitting simultaneously by a contact which opens when the TRC relay is actuated.

Auxiliary Equipment.

Recording and Logging Equipment.

The survey of various police and fire departments revealed that a need existed for sound recording equipment of a special sort. The principal new feature required was that the recorder should provide for imprinting a time and date stamp directly on the tape so that any recorded message could be identified with a particular stamp. Another objective was to obtain as much recording time as possible from a single roll of tape consistent with good speech quality. In this way, recording interruptions due to re-loading operations could be minimized, and tape as well as storage space conserved. To achieve this end, tape speed could be reduced and voice operation of the mechanism or operation by closure of remote contacts incorporated. Finally, of course, operation should be extremely simple and reliable and require very little maintenance.

To meet these requirements, Westrex Corporation developed the Signal Actuated Logger and Reproducer (SALAR) which consists basically of a

tape recorder capable of recording from one to four separate channels of information simultaneously on quarter-inch magnetic tape and of reproducing any of these channels. A block diagram of the recorder is shown in Figure 9. Up to 25- $\frac{1}{2}$ hours of continuous or intermittent recording can be accommodated on a single 10- $\frac{1}{2}$ inch roll of half-mil base tape at the recommended tape speed of 5/8 inch per second. Good speech quality is maintained at this slow speed by using reproducing heads with 0.1-mil gaps and by employing suitable pre- and post-equalization. A tape speed of 1 7/8 is optional and provides proportionally better frequency response but reduced recording time. With the voice-operation feature, starting delay is reduced by special design to only 10 milliseconds thus insuring that no intelligence is lost due to clipping of initial syllables.

To provide the time and date logging feature, a commercially-available time clock has been provided which stamps the magnetic tape automatically at the end of a message in the event that the machine is operating under voice signal control or alternatively at intervals of one minute if the machine is being operated continuously.

To assure that no information is lost in the event of tape run-out, or during reloading operations, two SALAR tape decks can be operated in conjunction with an automatic transfer feature which switches recording operations to the standby deck. Alternatively, the standby unit can be used to record up to four "secondary" channels until the automatic transfer occurs. Of course, additional decks can be used to record any number of channels.

The Function Selector shown in the lower part of block diagram of Figure 9 determines the operating modes. In the Manual position, continuous tape motion and recording are provided, whereas in the Automatic position, the tape is driven and recording takes place only when the unit is activated by voice operation or by remote control. With the selector in the Reproduce position, the bias oscillator is disabled so that recording does not take place, but any recorded information can be played back. In the Scan position, the drive puck is retracted, and the Tape Scan Powerstat is engaged to control the torque motors for fast forward or reverse tape motion enabling a particular recording to be located with minimum delay.

For clarity of presentation in the simplified schematic, only one of the four possible channels is shown. All of the electronic assemblies incorporated are of the transistorized type and are of etched circuit-board construction.

The basic version of the system comprises a single channel recorder-reproducer to which various optional accessories may be added.

The recording input circuit is designed to bridge across a 600-ohm line when the level is between -20 and +10 dbm and a front panel gain control is provided to accommodate this range of input level. For those applications where the input level is subject to considerable variation, an accessory compressor amplifier module is available which is basically similar to that described for the compressor amplifier for the console transmission system. The output level of the recording amplifier is maintained essentially constant for variations in input level over a 30 db range.

A second optional accessory is the voice operation amplifier, only one of which is required for each system for any number of channels one through four. On receipt of the audio signal, the output of this amplifier actuates the puck solenoid to engage the puck with the film and also results in connection of power to the feed and take-up torque motors. Cessation of the signal results in actuation of the stamping mechanism.

Separate reproducing amplifiers are provided for each channel and the outputs are available on the front panel at a nominal level of zero dbm for headset monitor or alternatively may be connected to remotely located power amplifiers for loudspeaker monitor.

Volume indicator meters are provided for each of the four channels and a push-button switch permits measurement of either the recording input or reproducing output levels. Circuits are available for connection to volume indicators for remote indication.

The input level control, recording amplifier, bias rejection filter, reproducing amplifier, reproducing level control, volume indicator meter and optional compression amplifier for each channel are all mounted in a single plug-in type module and modules are added as the number of channels is increased.

As mentioned previously, two such recorders can be employed together for four channel recording and a further optional accessory transfer panel permits automatic transfer of the recording circuits between the two units when the film on the first unit has been completely exhausted. This permits approximately 51 hours of recording time without the necessity of reloading.

Surveillance Systems.

In police department installations, various surveillance facilities may be required. Microphones may be concealed in interrogation rooms to provide means for monitoring and perhaps recording conversations held there. Cell blocks for prisoners might be similarly equipped to obtain information helpful in police work. By using voice-operated recording systems described elsewhere in this paper, round-the-clock surveillance of a number of such areas could be

automatically maintained. In smaller police stations, where personnel cannot be spared for guard duty, microphones could be strategically located in various widely-separated areas and connected to central equipment for continuous or intermittent monitoring.

For these surveillance facilities, commercially available microphones are used in conjunction with the specially-developed microphone and compressor amplifier units which are fully compatible with the other equipment provided.

Miscellaneous Specialized Equipment. Among other useful equipment in the final stages of development for police work is an adaptation of a Facsimile System. The U.S.A. Signal, Research and Development Laboratories and the Facsimile Section of Westrex Corporation have developed a facsimile system for transmitting pictures taken by a Polaroid Camera and recording the pictures at the receiving end on a Polaroid film. Within a short time, this equipment will be made available for civilian purposes.

The Military are using this equipment for several services. Its first application was radar meteorology. The radar scopes showing storm conditions were photographed with a Polaroid camera. Within a minute, the Polaroid picture was dropped into a slot and transmitted to the receiving station, with no more effort on the part of the transmitting operator than to press a button after dropping the picture in the slot. At the receiving station, the pictures were recorded automatically without any effort on the part of the receiving operator other than to remove the film in the same manner as he would remove film from a Polaroid camera.

The recordings may be in the form of a positive print or on transparent film. The transparent film recording enables the receiving station to project the picture on a screen by use of a conventional slide projector.

The transmission circuit between the transmitter and receiver may be a telephone line or a radio circuit that operates in the voice-frequency range. The signal is normally a 2400-cycle amplitude-modulated signal. Random noise in the transmission circuits seldom interferes with the readability of the received picture. Periodical noise, however, will degrade the received copy. However, in most cases, the result can be properly interpreted.

The type of modulation used in a radio link is not important. It may be AM, FM, pulse code modulated, etc.

The transmitted copy is scanned in the flat through an optical system employing an oscillating mirror. The recording light beam is also positioned by an oscillating mirror.

The scan definition is 200 lines per inch. The time of transmission for a Polaroid picture

having a rated size of 3 1/4 x 4 1/4 inches is 3 1/2 minutes.

One possible application for police work would be for the relaying to the central station from a patrol car the photograph or fingerprints of a suspect as an aid to identification, thus saving time and effort and perhaps avoiding lawsuits for false arrest which might arise if a person were taken into custody.

Another type of equipment which is available for police department application is a unit which is used to safeguard the privacy of conversations over radio-telephone and commercial land-line circuits. The unit is basically a speech inverter which passes a band of 250 to 2750 cps, inverts it about a mid-band frequency of 1500 cps and provides facilities for the transmit

and receive paths of a four wire circuit. The equipment consists of four parts, namely: two identical inverters for the transmit and receive paths, a common oscillator and power supply.

Conclusion

It is believed that the review of requirements of various police and fire departments has resulted in the design of new items which will find application in both large and small installations. It is considered further that the various items so developed are a practical solution to the various problems which are encountered in this type of installation and that considerable contributions can be made to the efficiency of operation of a system by their use.

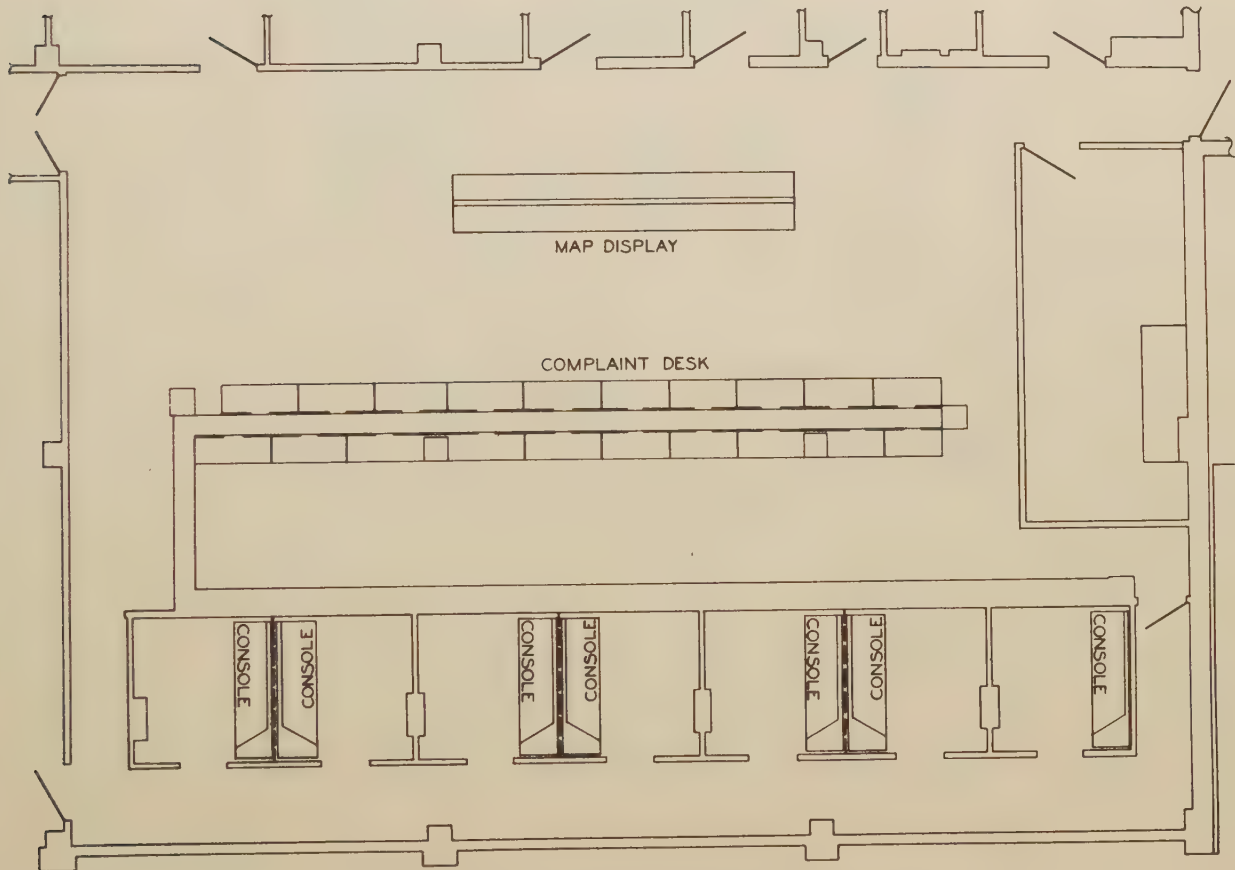


Fig. 1. Floor plan of a typical large command center.

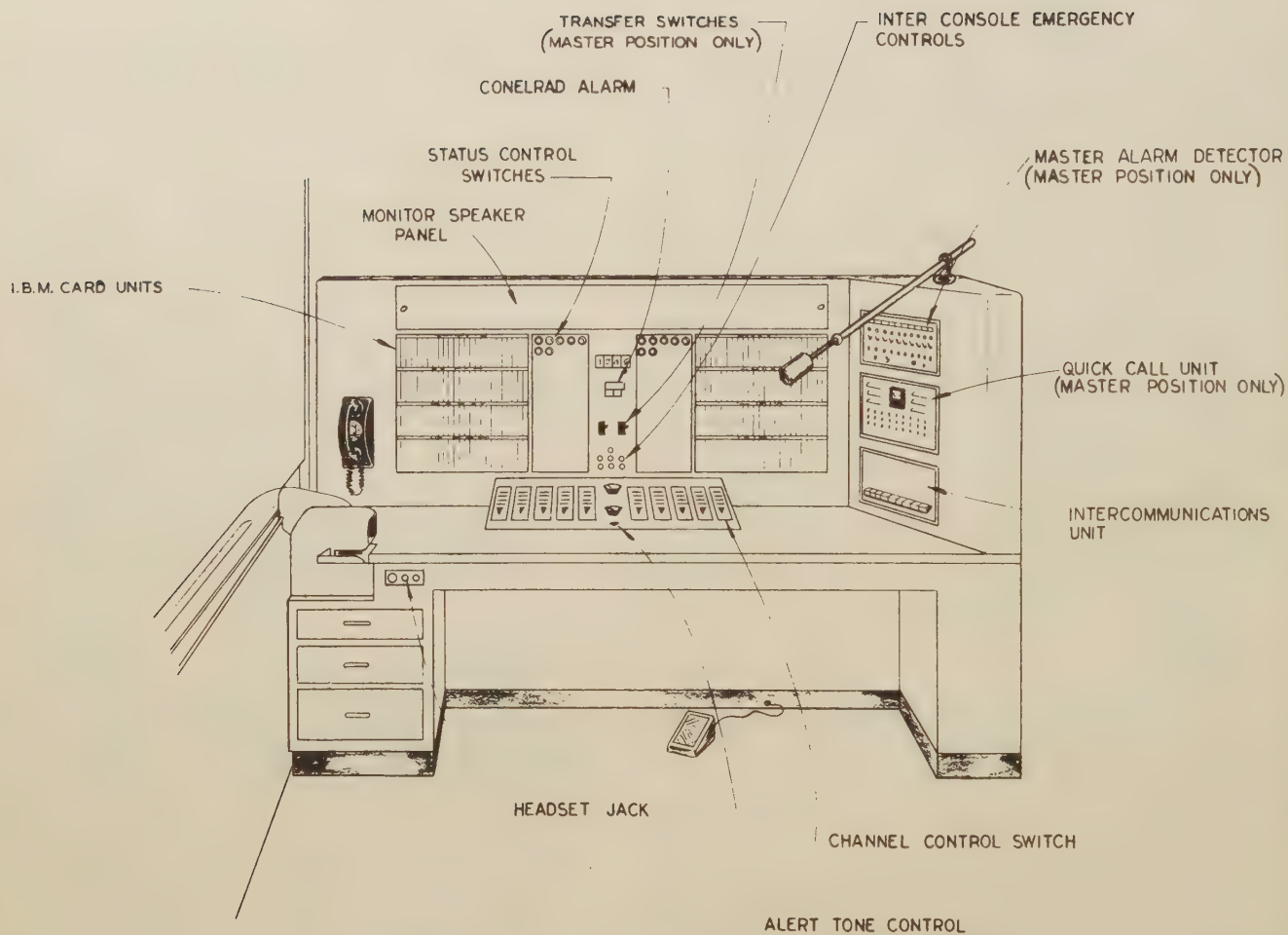


Fig. 2. Dispatch console for large command center.

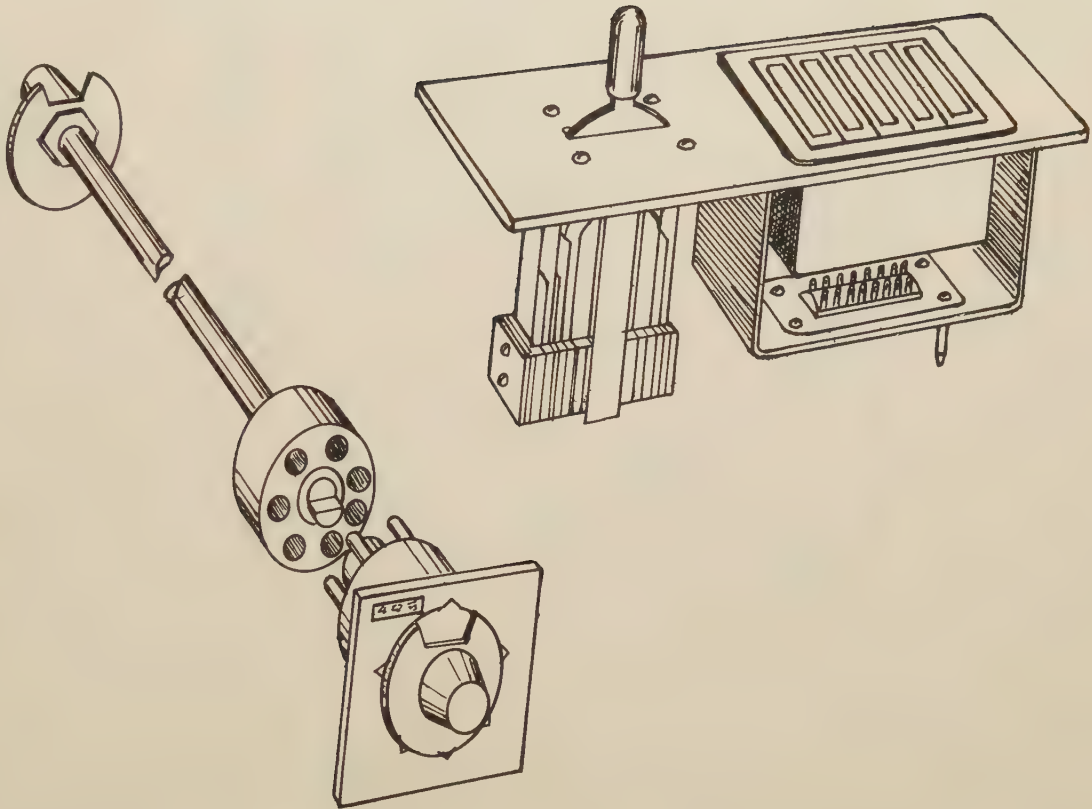


Fig. 3. Plug-in modules: Rotary status Switch and transmitter-control key assembly.

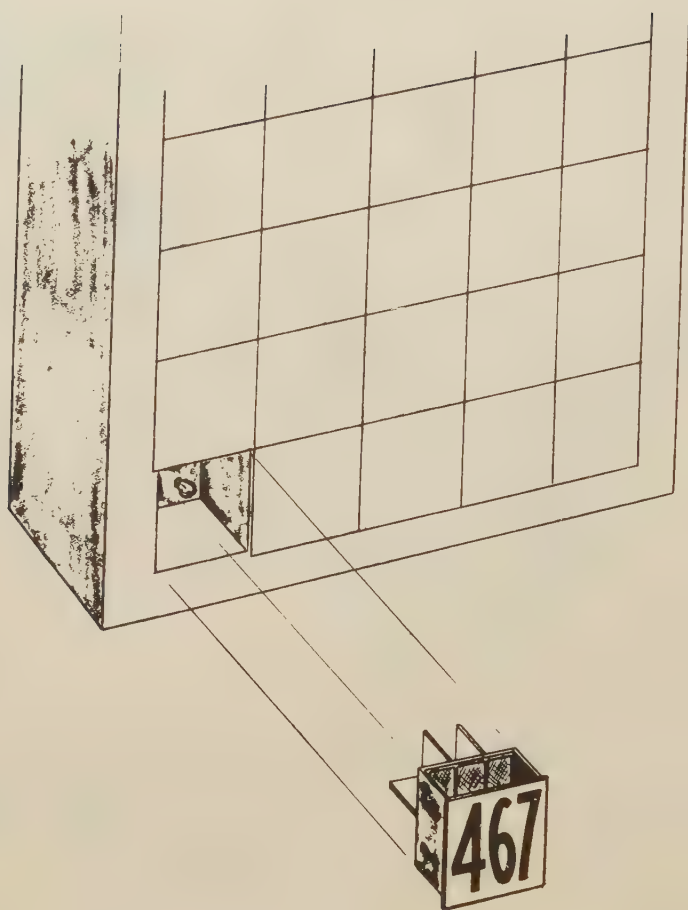


Fig. 4. Status-indicator cabinet.

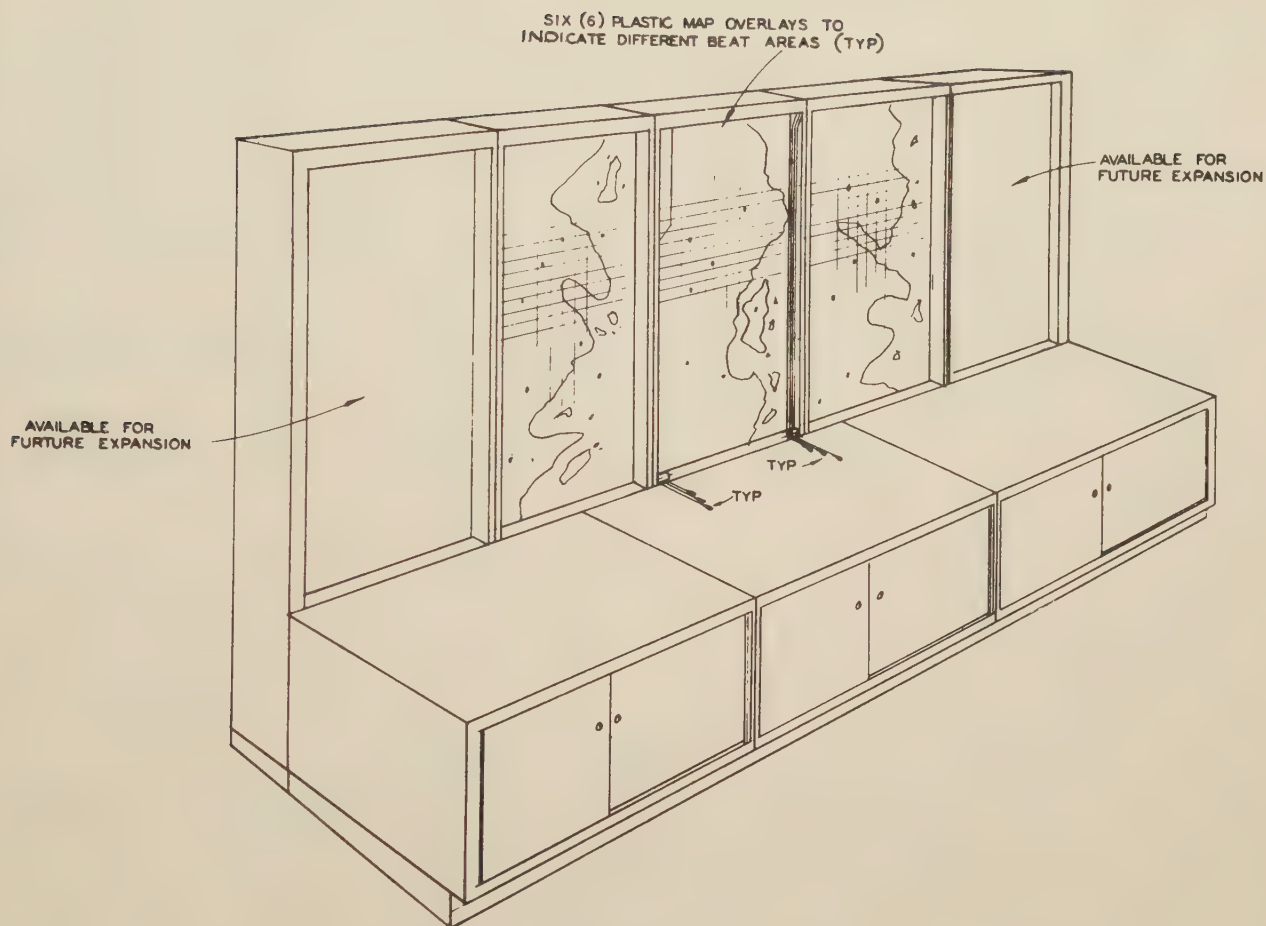


Fig. 5. Map display and storage cabinet.

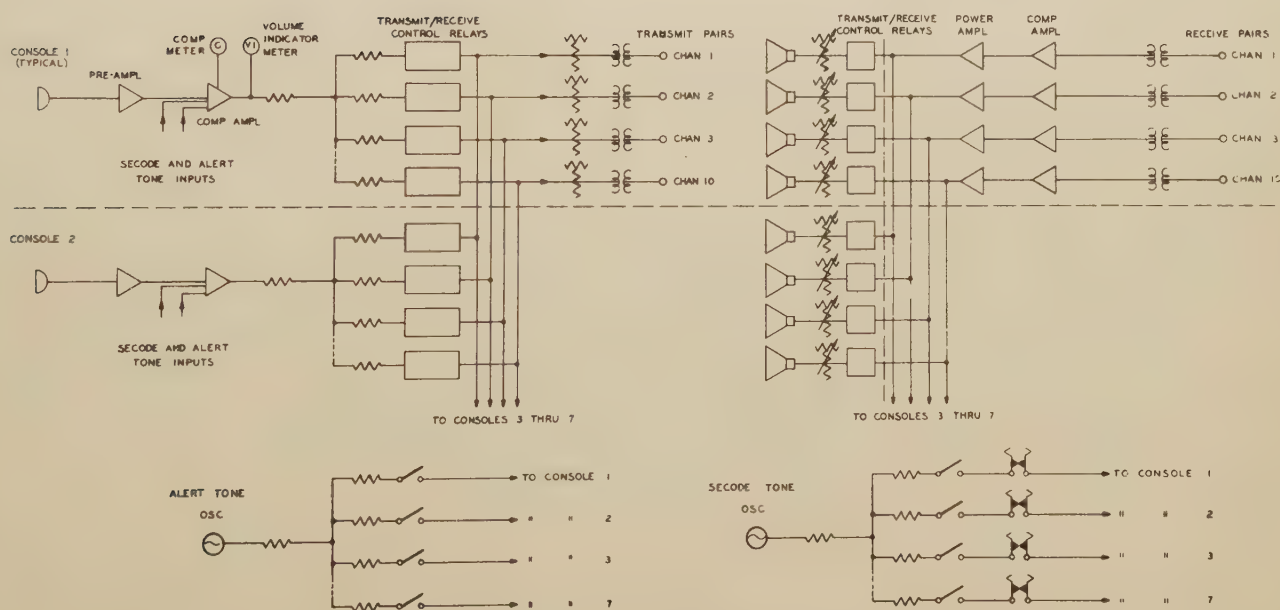


Fig. 6. Simplified schematic of transmission-receiving circuitry.

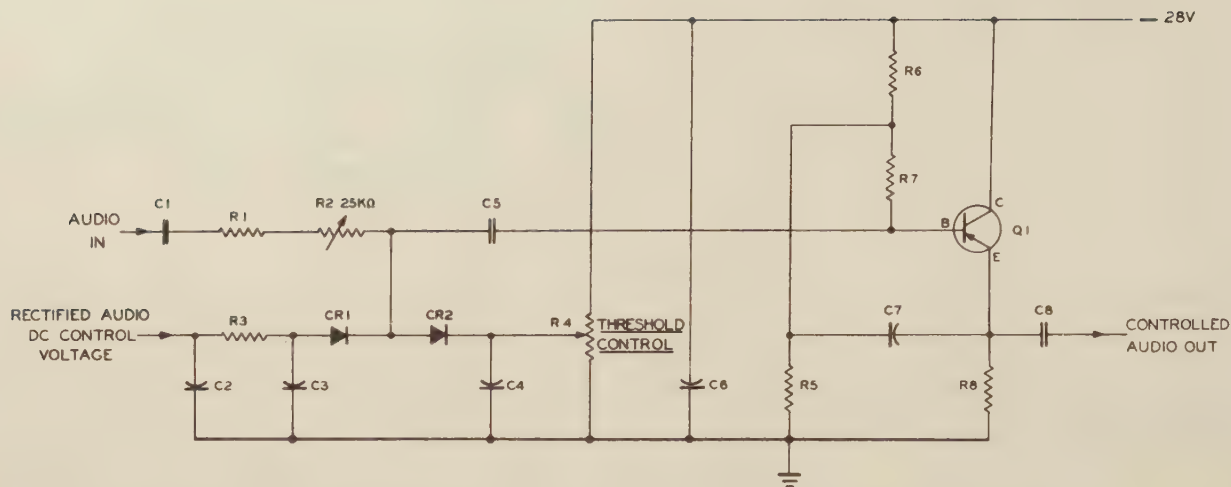


Fig. 7. Simplified schematic of compressor amplifier control circuit.

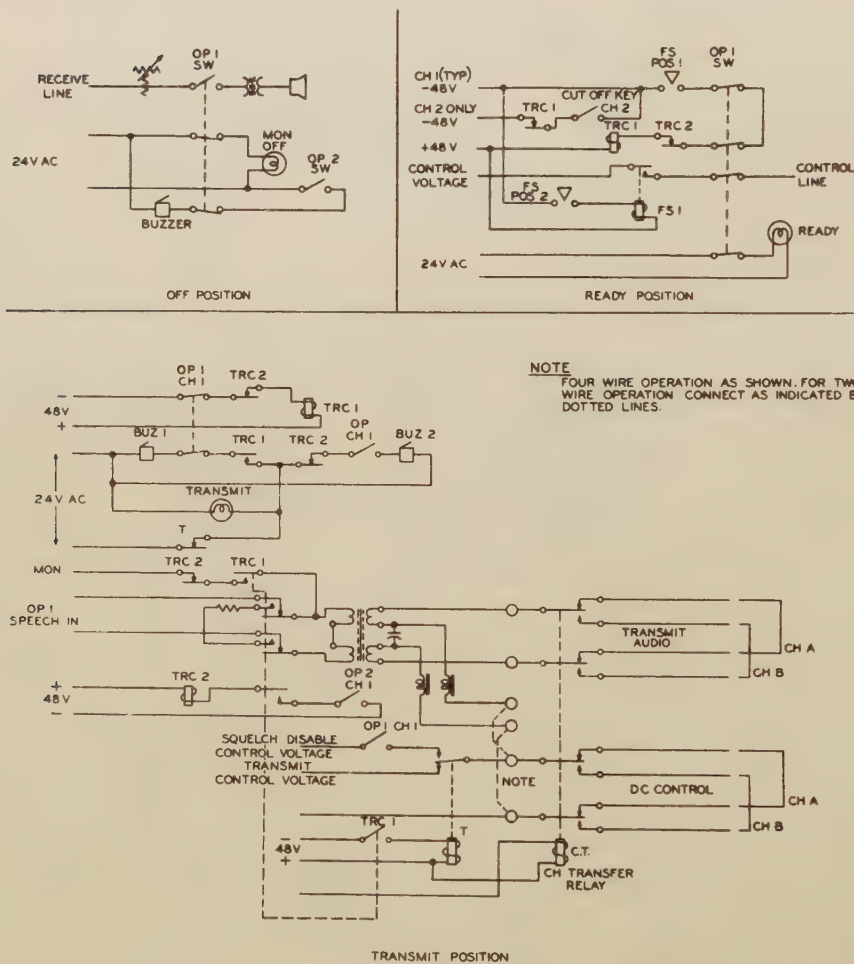


Fig. 8. Simplified schematic of switching circuitry.

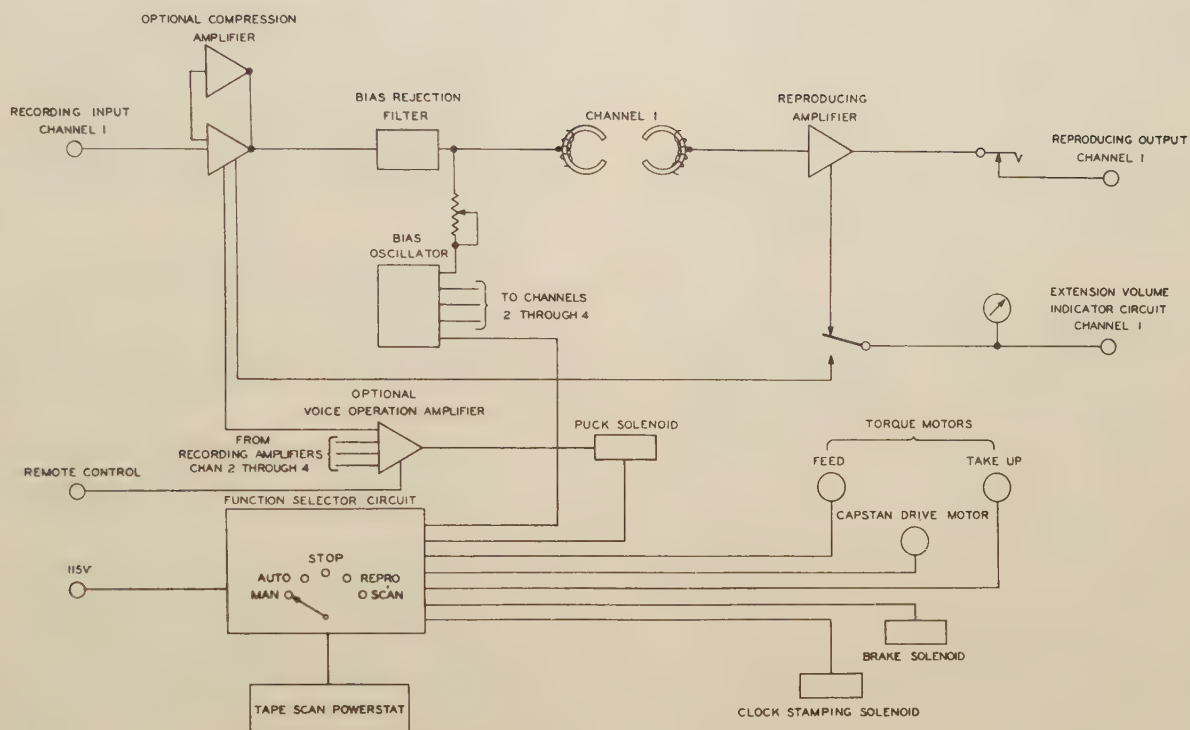


Fig. 9. Schematic block diagram of signal actuated logger and reproducer.

SYSTEM CONCEPTS FOR ADDRESS COMMUNICATION SYSTEMS

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Our Laboratory has been investigating the concept of address communication systems in a form which we call random access-discrete address. In the literature this is also referred to as non-synchronous or asynchronous communication.¹

The basic concept exploits the random character of speech so that some talkers are transmitting during quiescent periods of others. It is therefore a voice communication system although not necessarily limited to voice.

As we are considering the system many transmitters radiate signals more or less simultaneously, and each receiver must pick out of this mixture of signals those signals which are intended for him. Thus, we use the term random access. In order to accomplish this, each transmitter addresses each and every signal element which it transmits with the code of the intended receiver so that the receiver may select the coded signal addressed to it. Hence, the term discrete address.

A Random Access-Discrete Address System

A simplified version of a possible equipment is shown in Figure 1, which illustrates some of the features in the transmitter and receiver. The voice signal or supervisory signalling tones enters the pulse modulator where it is sampled and converted into a train of pulses which convey the intelligence. This might, for example, be a pulse position modulator or a pulse frequency modulator. Fixing attention on a single one of these pulses in the pulse train leaving the pulse modulator, it passes through a selected set of delay lines and these are tapped in a selected way and create thereby a group of pulses delayed with respect to each other and at different sub-carrier frequencies. The particular selection is chosen to constitute the code of the receiver being contacted. These pulses are then applied to an rf transmitter and radiated at the same rf frequency.

In the converse manner at the receiver, those pulses which are intended for the particular receiver will pass through the proper filters and delay lines in such a way as to arrive at the coincidence gate simultaneously. In this event, they will permit a pulse to leave the coincidence gate and either actuate a supervisory signal detector or pass through the pulse

modulator to produce the speech.

There are many variations possible to implement the concept and yet retain the basic features. There are also many techniques which can be applied to improve the specific operation of the system. Two techniques of major importance are blanking and correlation to reduce or eliminate errors caused by chance arrivals from two or more transmitters producing an apparent true signal.

Certain equipment design and operational advantages are evident in such a system. The system does not depend upon any synchronism either between the transmitter and the distant receiver or between the transmitter and its local receiver. This feature eliminates the need for high timing accuracies, a centralized timing system and the problems of finite transit time. Secondly, all transmitters in the system are transmitting at one rf carrier frequency, and this being a very wide band system because of the necessary short pulses, the tuning problem is avoided and the frequency assignment problem is greatly reduced. Thus, equipment to implement such a system appears to require less accuracy in the component tolerances than narrow band systems would require. Another feature is that we can have more than one code assignment to a given receiver and with little added equipment, be able to detect on each. One of these codes might be the address of this receiver only and hence be providing the equivalent of switched private line service. The other code might have an address common to a number of receivers and hence have the characteristics of the net mode. And a final point to notice for later reference is the tone signalling and tone detecting devices which are important requirements for proper system operation.

The points which have been mentioned are characteristics which have appeal even though there are problem areas to consider for a practical system. Hence, in addition to the characteristics of the basic system, consideration has been given to what happens when the system is extended to a large number of users and what can be done when distances exceed propagation range in the users equipment.

System Communication Areas

If the propagation characteristics are such that a receiver can receive signals generated

within the shaded area of Figure 2, the system will perform satisfactorily if the number of transmitters on the air within the shaded area does not exceed the capacity of the receiver. On the other hand, users of the system employing the same radio frequency and different but similar codes, might occupy the larger area. This does not mean of course, that this particular receiver could receive at this moment from all of these users but the system as a whole would contain more users than it normally would be able to support if it were not for the limitations of signal propagation. There are cases where such a characteristic is fruitful, since often the community of interest is limited to those who are within reasonable proximity.

In a mobile situation in which some users such as B might be leaving the area and other users such as C might be entering the area, the system will continue to function without saturation as long as the use density remains below the design level. In this case, community of interest is served whenever the propagation permits communication between the users. The advantage of not requiring the user to change frequencies or otherwise affect the operations as he moves from one area to another is more or less obvious.

The fact that this is a wide band system raises the question of the spectrum utilization of such a system and this is a very difficult point to evaluate because it depends so much on the point of view. Assuming a satisfactorily operating system such as shown in the figure, and a system requiring a total spectrum of six megacycles, any one of the following spectrum utilization values might be obtained. Suppose that in the shaded area there were 30 individuals actually talking, we would then require 200 kilocycles of spectrum per talker. If there were 300 users in this shaded area that would mean that there were at least 60 off hook conditions for these 30 talkers, which corresponds to everyone communicating 12 minutes out of the hour, which is a very heavy loading factor. From a frequency assignment point of view it can also be argued that these 300 users were being given substantial communication capability for 6 megacycles which then is 20 kilocycles frequency utilization per user. Finally, if there were 3000 users in the total area, operating within their individual areas of interest and using the same six megacycles, the spectrum utilization is 2 kc's per user of the total system. Thus the frequency utilization can be evaluated in a number of different ways to produce a good or bad evaluation.

Transponder

Turning to the capabilities of expanding the coverage beyond the limits of propagation of an individual transmitter, the range of coverage

can be extended through the use of a transponder in a variety of ways. One of these is illustrated in Figure 3, in which user #1, not being able to reach user #2 directly at the normal frequency F_1 , shifts his transmitter frequency to F_2 . The transponder receives all frequencies at F_2 and retransmits on F_1 , without demodulation or detection of the address codes. The receiver at user #2 then receives the codes which are addressed to him at his normal frequency, F_1 and responds to user #1 using the same technique through the transponder. Thus, increased range of coverage can be handled in rather a simple manner without adding the additional functions of local switching centers.

The service being provided to the users is very similar to that provided by local switching facilities of a telephone exchange. However, since the telephone exchange is absent some parts of the service necessary to make a complete system operate satisfactorily which had been supplied by the local exchange, must now be incorporated in the equipment. Since in Figure 3 it had been assumed that the two users were out of range, listening at the proper address would not indicate whether the person to be called was or was not in a busy condition. It is therefore felt that to initiate a call it would be necessary to address the called party with a supervisory signal, perhaps in the audio range, using the called party's address as in speech transmission, much as you would transmit the ringing signal on a telephone circuit. The receiver must then have the capability of indicating to the calling party either that the called party was ready to receive a call (phone off the switch hook condition) or that the called party's line was busy. These are the supervisory conditions incorporated in the tone signaling boxes of Figure 1. It would also be desirable in the case of using a transponder to indicate through supervisory signals those calls which have arrived via the transponder so that the called party can respond via F_2 , the radio frequency for transponder operation.

Double Transponder

Figure 4 illustrates how one can apply the transponder principle and provide simple trunking between two areas, such as between two towns or two military complexes. In this case, the output of the receiver containing all the address codes which it receives is patched, on a wideband basis, into a relay transmitter which is then received at the distant location and re-radiated at the radio frequency F_1 for users of the system near this distant station. Replies would return via a similar path. One thing to observe in this mode of operation is that from the user point of view no new capability needs to be incorporated into the user equipment. All receptions still remain on the same rf frequency

F₁ whether it comes from a transponder or a user close enough to be contacted directly and all transmissions which cannot be contacted directly are radiated on the same rf frequency F₂.

Triple Transponder

As long as the capacity of the system requiring extended range coverage is not exceeded, more than one remote relay transponder station can be employed either on parallel legs as shown in Figure 5 or these can even be connected in tandem. The reception of the signals from a particular remote transmitter provide sufficient information for a given station to radiate and to even re-transmit to another station without ambiguity or without introducing "ring-around-the-rosy" loading of the system. As such a system is expanded it will eventually reach a saturation point in which the trunking circuits will be over-loaded and it would be profitable to introduce switching.

Address-Switching Center

Other than the normal functions of reception and transmission a switching center, Figure 6, would then be required to have a bank of address detectors which could be addressed in the same manner as one would address a remote party except that the user in this case would be required to find an idle address detector just as the operator is required to find an idle trunk. The output of such an address detector then would be switched through switching equipment and into the appropriate re-addressing device for transmission through the proper channel of the switched portion of the system on a random access-discrete address basis. Such a system would require intelligence to be built into the switching center, similar to that which is now provided for long distance dial switching and address detectors and re-addressing equipment to replace the functions now performed by the multiplexing equipment in multichannel trunk switching centers. It appears reasonable to assume that the complexity of such a switching center would not be too different from that which is now employed. In fact, it might result in less equipment than now is required although such a claim cannot be made at this time.

Net-Switched-Address Concepts

Some of the system concepts of a new form - address type - of communication system design

have been described. When we wish to converse with an individual in the net mode the user selects a proper frequency for calling his party. When we use the switched mode two or more circuits are connected at a switching center to provide the connectivity required. Address systems are distinguished from both of these by having the character of the signal in addition to carrying the message information, also contain the information which distinguishes the called party. Ordinarily, this requires more bandwidth than is required by transmitting the intelligence information only. The interesting thing about it is that when viewed on a total system basis this excess bandwidth is compensated for by the idle periods between conversations.

It is possible to provide switched type characteristics (such as by a single frequency assignment for a single user) in the net mode. It is also possible to provide the net form in a switched mode, for example, by a conference circuit. Generally speaking, any given communication system problem is more suitable for one or the other of these. However, in the address system, it is convenient to provide both types of service simultaneously with minimum additional equipment. Thus there now is a new domain for system engineering of communication systems.

In closing, the addressing technique which was described in Figure 1 is only one of a number of possible ways of addressing. Any reasonable method of incorporating in the signal sufficient distinguishing characteristics to provide a large number of addresses without at the same time introducing an exorbitant number of signals is suitable for this technique. Work in these areas has been increasing of late and because of the system characteristics inherent in the concepts we can expect to have an increasing amount of work accomplished in these areas.

¹Taylor, J. E., "Asynchronous Multiplexing", Communication and Electronics, January 1960.

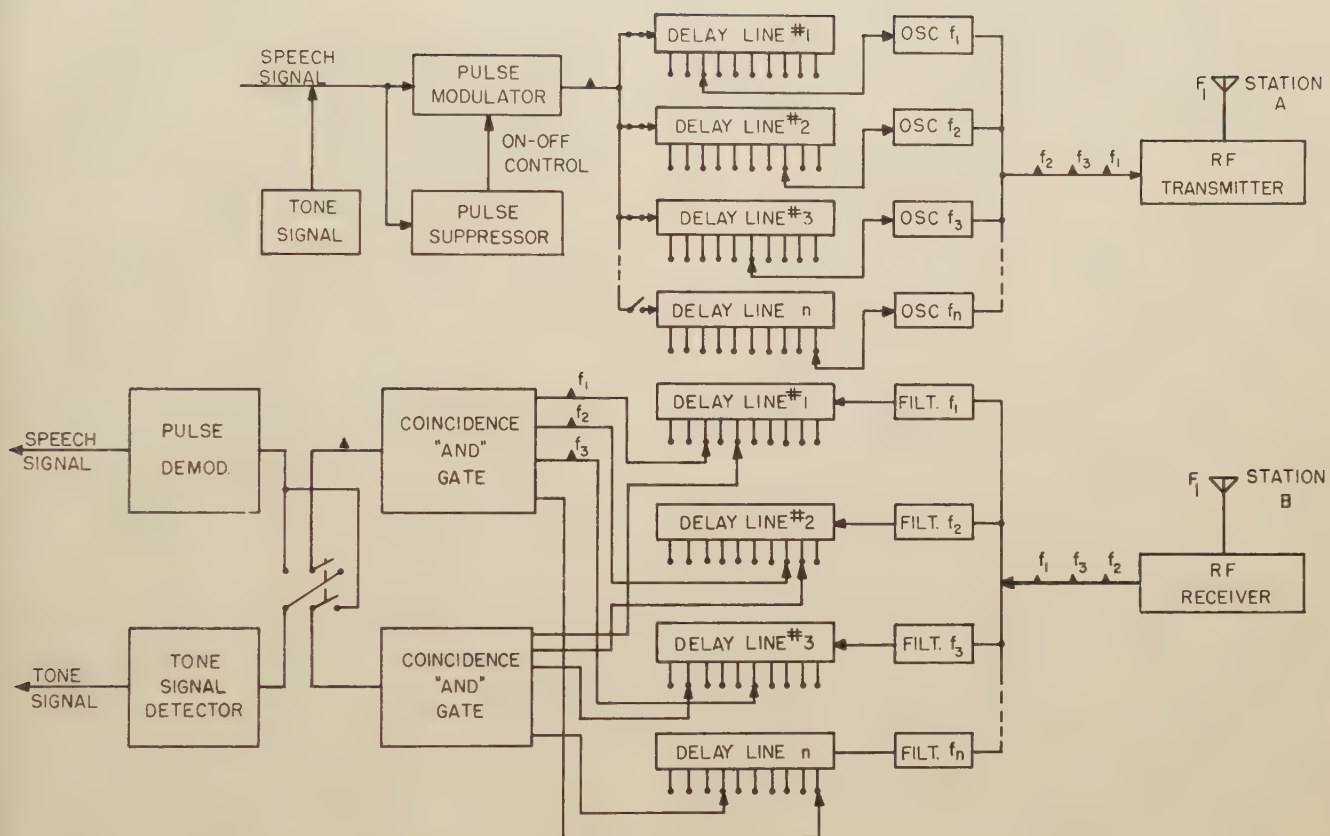


Fig. 1. A random access-discrete address system.

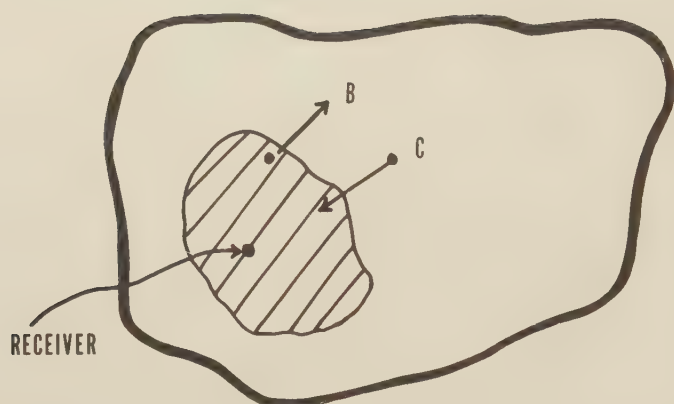


Fig. 2. System communication areas.

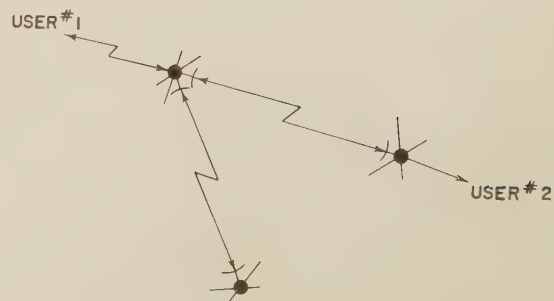


Fig. 5. Triple transponder.

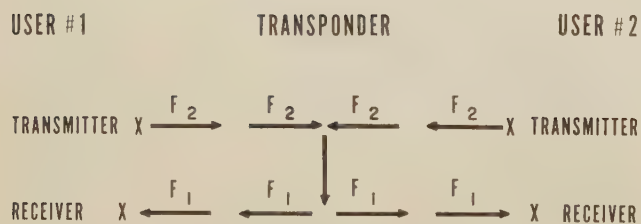


Fig. 3. Transponder.

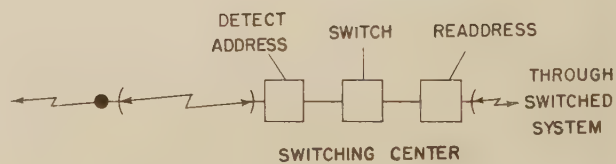


Fig. 6. Address-switching center.

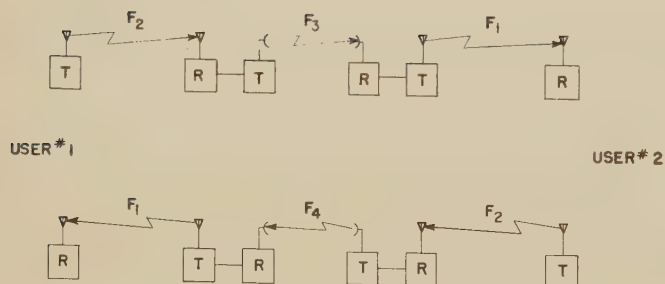


Fig. 4. Double transponder.

PUSHBUTTON DIAL MOBILE RADIOTELEPHONE AN ADVANCED CONCEPT IN COMMON CARRIER MOBILE SERVICE

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Summary

A real need has developed for common-carrier mobile radiotelephone service which can be connected directly into dial telephone exchanges. Many exchanges are now unattended and the mobile subscriber must be able to dial his calls and receive calls in the same manner as with his office or home dial telephone.

The transmission of the dialing information from a moving vehicle into the exchange, with the accuracy required to meet telephone standards, has presented an interesting challenge in view of various radio propagation characteristics often experienced, particularly from weak signal or fringe areas.

The pushbutton type of dialing described in this paper provides a digital signalling system employing tones of sufficient duration to ride over the numerous signal drop-outs sometimes lasting more than 100 milliseconds. Extensive field experience has proven this approach to be highly reliable and consistent in dialing from a mobile unit. A unique application of counter tubes and transistor switching circuits is employed in the conversion of the tones into dial pulses at the base station terminal. The use of pushbuttons has been found to be very convenient for mobile service.

A number of the various features incorporated in mobile dial systems are described in this paper.

Introduction

A pushbutton "dial" mobile radiotelephone system has been developed which provides a reliable method of connecting common-carrier mobile units into dial telephone exchanges. Dialing by means of pushbuttons has been found to have distinct advantages in mobile radio service, including an ease of operation, particularly while the vehicle is in motion. It provides a high degree of reliability in the method of tone signalling employed from mobile to the exchange base station.

A number of telephone operating companies are now offering the pushbutton radio-

telephone service to their subscribers, which gives complete dial telephone service to and from automobiles and trucks. Telephone calls, either local or long distance, can be made and received in much the same manner as a regular dial telephone in the home or office.

Overall System

Radio terminal equipment at the dial exchange ties the mobile radio system into the standard telephone circuits of the exchange. The connections into the exchange can be via either subscriber lines or two-way dial trunk circuits, depending upon the type of ringing available in the exchange, and the type of selective calling employed in "ringing" the mobile subscribers. In most cases no modifications are required in the exchange equipment.

Figure 1 shows the over-all system with the control terminal equipment located adjacent to the dial exchange equipment. The control group includes the line coupling circuits to and from the radio base station, the compression amplifier for the transmit audio, the tone generators used in selective calling of the mobiles, the tone decoders for converting the pushbutton tones received from the mobiles into regular dial pulses, and various control circuits to provide the normal telephone supervisory signals into the exchange.

The radio base station consists of a 50-watt FM transmitter and a sensitive receiver, as commonly used in regular mobile systems operating in the 150-174 mc band, arranged for duplex operation and for prolonged duty cycling required in telephone service. The transmitter is under control of the exchange and is automatically turned on when the system is in use. The base station is often located remote from the exchange, preferably on high ground or on top of a tall building to give desired coverage. A 4-wire audio circuit is used to connect to the control equipment usually mounted within the exchange.

Any local "land" subscribers of the exchange can call a mobile subscriber by dialing the directory number of the desired mobile unit in the regular manner. Likewise the mobile

subscribers can "dial" any land line subscriber. Long distance calls can be handled to and from the radio system. The toll calls can be completed either through operators, or if the system includes mobile identification circuitry, through the automatic toll ticketing.

The mobile subscriber's equipment is shown in Figure 2. A standard 25-watt 2-way radio set mounts in the rear trunk compartment of the car, or in some convenient location of a truck. The pushbutton control head usually mounts on or just under the dashboard of the vehicle. It controls the various tone signals transmitted from the mobile unit. On the front are 10 illuminated pushbuttons, one for each digit of a standard telephone dial. Also located on the front are the supervisory lamps for indicating an incoming call, a channel busy condition, and the turn-on of the mobile transmitter. The illuminated panel and pushbuttons indicate that the radio set is turned on and in a standby condition.

A push-to-talk handset mounts adjacent to the control head and normally rests in a hang-up cup which includes a hookswitch arrangement.

Factors Involving Tone Transmission

Audio tones within 1300 to 2700 cps are used to convey the supervisory signals from the mobile unit such as line seizure, dialing, and disconnect.

The characteristics of radio wave propagation to and from moving vehicles have to be considered in any mobile system involving tones. Due to multi-path reflections, especially in fringe or weak signal areas, standing wave conditions are often experienced. These conditions may result in gaps or holes in the received signal level.

Figure 3 shows a reproduction of an actual recording of typical levels of a 150 mc signal received from a vehicle moving at about 20 miles per hour in an urban neighborhood near the fringe area of the talk-back range. It can be seen that gaps of over 100 milliseconds occur. These holes may be unnoticed during voice transmissions, but can seriously affect tone signals, especially those used for digital information such as dial pulses, the standard rate of which is 10 per second or 100 milliseconds per pulse.

In the pushbutton dialing system described

here each tone is automatically timed for a duration of 400 milliseconds which is sufficiently long to ride over the gaps and thus assure reliable registration by the decoder circuits in the control equipment at the exchange.

Pushbutton Dialing

A sequential two-out-of-four tone arrangement is used for the dialing tones. The tone selections are made by the pushbuttons as shown in Figure 4. Each tone is transmitted for a duration of 400 milliseconds as mentioned above. Immediately following the two dialing tones is an interdigit tone for each digit or each pushbutton depressed. This third tone is used at the exchange control group to trigger the conversion of the tones into digits. The pushbuttons tune the frequency of a common transistor oscillator circuit in the control head. Stabilized resonate networks are selected in accordance with the pushbutton depressed and a switching circuit timed at 400 milliseconds. The four dialing tones are between 1900 and 2500 cps. The same oscillator is also used for various supervisory tones, including identification, down to 1300 and up to 2700 cps. Since one tone is transmitted at a time, it can fully modulate the mobile transmitter and thus produce a maximum signal-to-noise ratio.

As soon as a button is depressed it becomes locked in the operated position by a solenoid which in turn is kept energized by the same timing circuit that controls the tones. The button remains depressed approximately two seconds or long enough to transmit the associated three tones and to convert them into the dial pulses at the exchange. The two-second duration allows 1.2 seconds for the transmission and up to 1.0 second for the conversion in case the digit is a zero or equivalent to 10 pulses. As soon as the button releases the next button can be depressed. Only one button at a time can be operated. Operation of the buttons has been found to be very convenient for mobile service.

Figure 5 is a block diagram of the dialing circuits. The depressing of a button energizes the locking solenoid and arranges the transistor oscillator for transmitting the two dialing tones plus the interdigit tone through the automatic timing circuits. The mobile transmitter is also keyed by the pushbutton.

At the base station end the tones are fed to the decoder circuits in the control group. The proper digit is registered in the digit counter tube. The interdigit tone is decoded and triggers the pulser counter tube, which in turn pulses the

dial pulsing relay. Contacts on the relay open and close the d-c closure on the line into the exchange in the same manner as a regular telephone dial.

Digit and Pulser Counter Tubes

Figure 6 shows the unique application of the cold cathode counter type tubes, commonly used in computer equipment, to dial pulsing circuits in the control group. Both tubes are the same, each having 10 cathodes, as either Baird-Atomic GS10C or Sylvania 6476. The digit tube is used to register the digit number dialed from a mobile subscriber. The pulser tube controls the dial pulsing relay at the conventional telephone dial rate of 10 pulses per second, with each pulse following the standard 60% break and 40% make. The cathodes have been numbered in accordance with their functional positions in the circuit operation. In idle condition, the glow in the digit tube is resting on cathode 0, and in the pulser tube the glow is on cathode 2.

Digit Registration. Upon receipt of two dialing tones, the associated decoder relays will operate. Through a combination of made relay contacts, a low potential is applied to the cathode corresponding to the digit dialed from the mobile subscribers. For example, assume the digit 7 button had been depressed in the mobile unit. Tones at frequencies of 1900 and 2100 cps are received. A low potential is momentarily applied to cathode 7. The glow is immediately transferred from cathode 0 to cathode 7. Current flows from cathode 7 through the base resistor of transistor Q6 to unbias the transistor and in turn operate the off-normal relay. The relay will remain operated until the glow returns back to cathode 0. The relay contacts have prepared a path for the eventual operation of the stop relay.

Dial Pulsing. Upon receipt of the inter-digit tone, the associated decoder relay energizes the 100 cps driver circuit for the pulser tube. The glow immediately starts to move from cathode 2 through the other cathodes in a left to right direction stepping at the rate of a cathode every 10 milliseconds. As soon as the glow has "hit" cathode 5, the transistor Q1 becomes unbiased and the operate coil of polar-type pulsing relay is energized. Contacts of the pulsing relay open the line closure into the exchange.

The 100 cps driver circuit will continue to move the glow through cathodes 6, 7, 8, 9 and 10. At cathode 10, the digit tube drive circuit

is momentarily triggered and steps the glow in the digit tube one cathode to the right, i.e. from 7 to 6. Therefore, it can be seen that for each rotation of the glow around the pulser tube, the glow in the digit tube steps one cathode.

The glow continues rotation in the pulser tube and "hits" cathode 1 which actually is adjacent to cathode 10 in a radial arrangement. With the glow on cathode 1, current flows through the base resistor of transistor Q2, unbiasing it and in turn energizing the release coil of the pulsing relay. The relay contacts now close the d-c loop into the exchange. The relay remains released as the glow continues through cathodes 2, 3, and 4.

Thus it can be seen how accurately the dial pulses are formed for sending into the exchange. The 100 cps driver steps the glow in the pulser tube at the rate of a cathode every 10 milliseconds, through the 10 cathodes for one complete dial pulse in a period of exactly 100 milliseconds, which is at the conventional dial rate of 10 pulses per second. During six of the cathodes the pulse is in its break portion for 60% of the time, and during four of the cathodes the pulse is in the make portion for 40% of the time.

As mentioned above, for each revolution of the glow in the pulser tube, the digit tube glow moves one cathode. Therefore, in the example of digit 7, the pulser tube glow has rotated seven times when the digit tube glow comes to cathode 0. At this point the transistor Q5 becomes unbiased and the stop relay operates. Contacts of the stop relay "ground" the 100 cps driver circuit to disable it. A second set of contacts also applies grounds to the home positions of both tubes. As the glow in the digit tube left cathode 1, the off-normal relay starts to release but being slow to release its contacts remain closed long enough for the stop relay to operate. When the off-normal relay finally does release, the stop relay also releases and the circuit is ready to receive the next dialing tones.

On receiving the tones for a digit 10 (or 0), the same procedure is followed, except that during the first rotation of the pulser tube the digit tube glow remains on cathode 0. At the end of the first rotation the digit tube glow moves to cathode 9, the off-normal relay operates, and then the procedure continues as previously described. Thus it can be seen that the pulser tube will have made 10 revolutions producing the required 10 dial pulses.

Supervisory Tones

The mobile control head also produces the various supervisory tones as shown in Figure 7.

Line Seizure

When the handset is lifted from the hook-switch to initiate a call, a combination of line seizure tones are transmitted. In addition to seizing the line or trunk in the dial exchange as an "off-hook" signal, these tones can be used for mobile identification. The first two tones identify the particular mobile subscriber and the third tone identifies the loop or line assigned to the subscriber. For example, if the system connects into the exchange via 10-party subscriber lines and up to three such lines for 30 mobile units, the loop tone will seize the proper line.

Figure 8 shows a functional diagram of the line seizure circuits and the fact that both an r-f carrier and the proper tones are required to seize the line. When the handset is lifted to initiate a call, operation of the hookswitch turns on the mobile transmitter for approximately two seconds and the prearranged seizure tones are sent to the base station. In the control equipment of the base station, the squelch slave relay operates and locks on a timed circuit. The associated loop tone relay also operates and through a control relay the assigned line is connected to the audio circuits of the base station. As soon as the line has been seized by the exchange, the base transmitter is turned on by the sleeve or control lead of the line and the dial tone is transmitted back to the calling mobile subscriber.

Disconnect

When the handset is restored at the end of the call, two tones, at the frequencies of 2300 and 2500 cps, are transmitted to release the locked-up squelch slave relay and to open the d-c loop closure into the exchange in same manner as an "on-hook" signal.

If the mobile subscriber has moved out of range of the base station during the call and thereby is unable to send in the disconnect tones, the timed lock-up circuit will eventually release the line. The timing interval is usually adjusted to approximately three minutes.

Ringling Out to Mobiles

As mentioned earlier, the type of ringling out to the mobile subscribers is related to method of connecting into the dial exchange, i.e. either via subscriber lines or by trunk circuits, and, if via subscriber lines, the type of ringling available in the exchange.

The control equipment can be arranged to accommodate a number of different types. The method described here is assuming that the radio system is connected into the exchange via 10-party subscriber lines and that the exchange has fully selective multi-frequency ringling, as commonly found in many exchanges, particularly in rural areas.

Figure 9 shows a block diagram of the electronic ringling circuits in the control equipment as connected into a 10-party subscriber line circuit. Assume the exchange has decimonic ringling, i.e. frequencies of 20, 30, 40, 50, and 60 cps, and that the ringling can be applied to either the L1 or L2 side of line. As soon as the connector equipment of the exchange has seized the line for an outgoing call, a battery condition has been placed on the "C" control or sleeve lead and the base transmitter is turned on. If the ringling is appearing on the L1 side of the line, the associated line sensing circuit energizes the L1 oscillator and its stable reed-controlled 170 cps frequency output is fed to a common balanced modulator. The ringling signal is also fed to the modulator. The resultant output after mixing will be two tones comprised of the lower and upper sidebands of 170 cps. Assume the ringling is 20 cps, then the two sidebands will be 150 and 190 cps. These two tones are transmitted simultaneously at the ringling interval. The ringling tone decoder in the called mobile is tuned to these two frequencies and will energize a bell as well as a call light. The bell follows the ringling intervals and the call light remains on. If the subscriber was out of the vehicle, the call light would indicate that he had received a call.

It can be seen that if the ringling signal had been another frequency or had been on the other side of the line, a different combination of ringling tones would have been transmitted. Thus any one of the ten mobiles assigned to the loop can be selectively called. Other loops connected into the system have their associated ring tone oscillators tuned to different frequencies and thus extend the selectively calling to any one of all mobiles in the system.

Mobile-to-Mobile Calls

On mobile-to-mobile calls the base station functions in the same manner as a mobile repeater station with the transmitter under control of the exchange.

With the system connected into the exchange via subscriber line circuits as described above, the calling mobile subscriber dials the called mobile party in the conventional manner. A short duration unbusy circuit is included in the control group to permit the exchange connector to seize and ring out on the line associated with the called mobile unit. In case both mobiles are on the same line, the reverting call arrangement of the exchange is followed in dialing the desired party. For such calls the standard procedure generally instructs the calling party to hang up following the dialing and await the answering of the called subscriber. An eleventh button labelled "PTY" is mounted on the front of the control head and when it is momentarily depressed, an "on-hook" signal is transmitted to the exchange, accomplishing the same function as the calling party hanging up. Therefore, in the mobile service the calling party can keep the handset in his hand and immediately talk as soon as the called mobile answers.

Remote Fixed Subscribers

In a number of the mobile dial systems, it has been found that a need exists for remote fixed station equipment, particularly at temporary locations such as resort areas and construction sites where permanent wire lines are uneconomical. Another common application has been in off-shore installations.

The remote fixed subscriber operates into the mobile system in same manner as a mobile subscriber. A telephone desk set, with a push-to-talk handset, mounts on top of the pushbutton control head and is connected together on a 4-wire audio basis.

The 25-watt radio set is powered from a 117 volt, 60-cycle source. A corner reflector-type antenna is generally used and often mounted on a 40-to 60-foot telephone pole just outside the building or residence.

Privacy of Conversation

All mobile and remote fixed subscribers except the one rung are normally locked out.

They are not able to listen in on the conversation nor are they able to turn on their transmitters. The In-Use light on the control head is on whenever the channel is in use.

There is provision, however, for a radiotelephone party to break in if he has an emergency call. A set procedure must be followed which involves removing the handset from the hookswitch, turning the power switch off for approximately four seconds, and then turning it back "on", and depressing the push-to-talk button on the handset. This will automatically break into the existing call and enable the party to request that an emergency call be given priority.

F. C. C. Regulations

Mobile dial systems are licensed under Federal Communications Commissions Rules and Regulations, Part 21 "Domestic Public Radio Services". Several special features have been incorporated in the system equipment to assure compliance with the regulations.

One is the emergency break-in described above, so that "messages concerning the safety of life or property which is in imminent danger shall be afforded priority over other messages".

Another feature is the automatic base station code identifier. At the end of each call, the identifier located in the control group keeps the base transmitter on for approximately five seconds and automatically transmits the call letters of the base station in Morse Code.

An important feature provides for the controlling of the base station by means of the exchange battery. So if the transmitter is found to be on when for some reason it should not be on, removal of the exchange battery at the control group will turn the remotely located transmitter off.

To comply with the Conelrad alert regulations, a Conelrad monitor receiver can be installed with the control group to prevent the base transmitter from being turned on during an alert. An automatic reset can be included with receiver to restore the base station to normal at the end of the alert.

Conclusion

The performance of the existing push-

button radiotelephone systems in various parts of the country has proven the reliability of the 400 millisecond tones and of the sequential 2-out-of-4 tone arrangement. Subscriber acceptance of pushbutton dialing has been received. The operation of pushbuttons in a car has been found to be convenient and simple. The ever increasing interest shown in mobile dial indicates that many more communities will soon be offering common-carrier service with push-button dialing.

Acknowledgment

The author wishes to acknowledge the efforts of all who contributed toward the development and initial installations of the push-button radiotelephone systems. In particular, credit is given to Messrs. Charles Willyard, William Wray, Tony Sakota, and Jack Swartz.

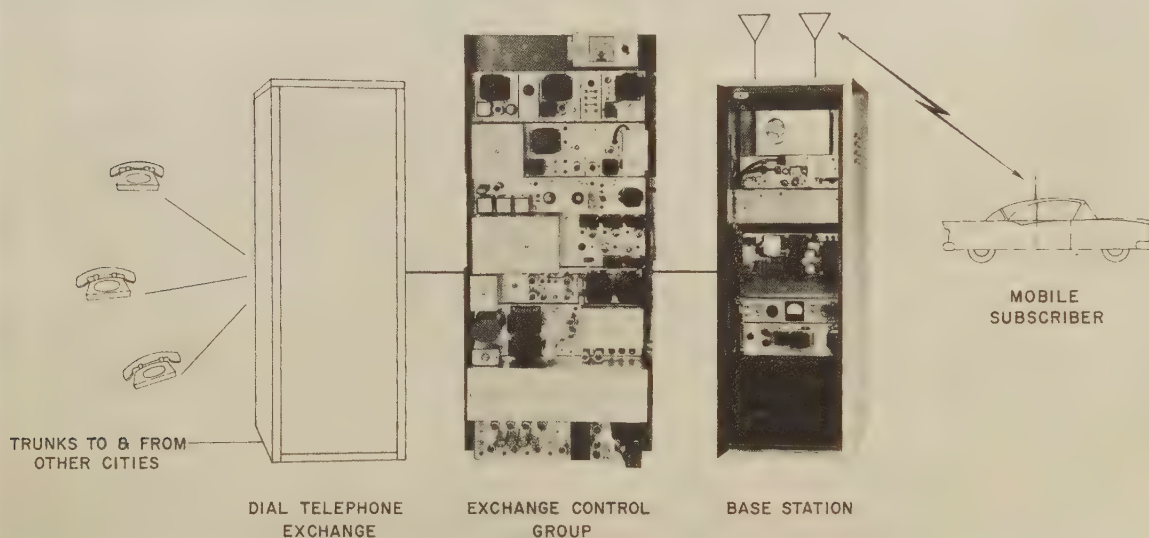


Fig. 1. Pushbutton "dial" mobile radiotelephone system.

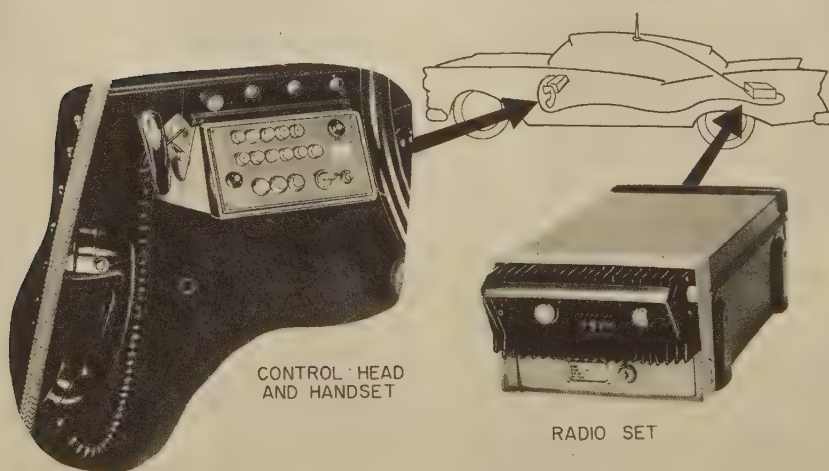


Fig. 2. Mobile subscriber's radiotelephone.

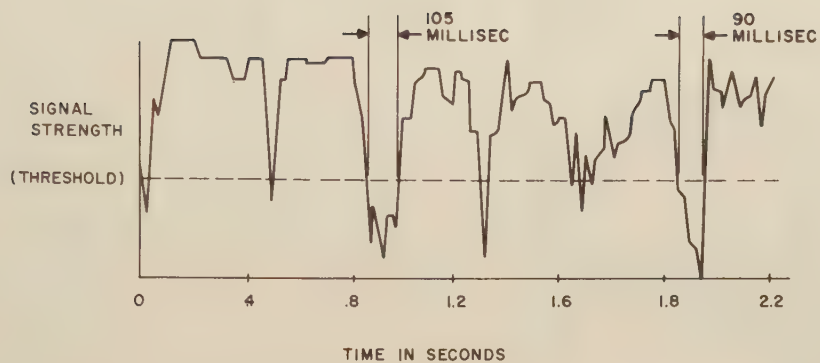


Fig. 3. 150-mc signal strength variations received from a vehicle moving at 20 mph.

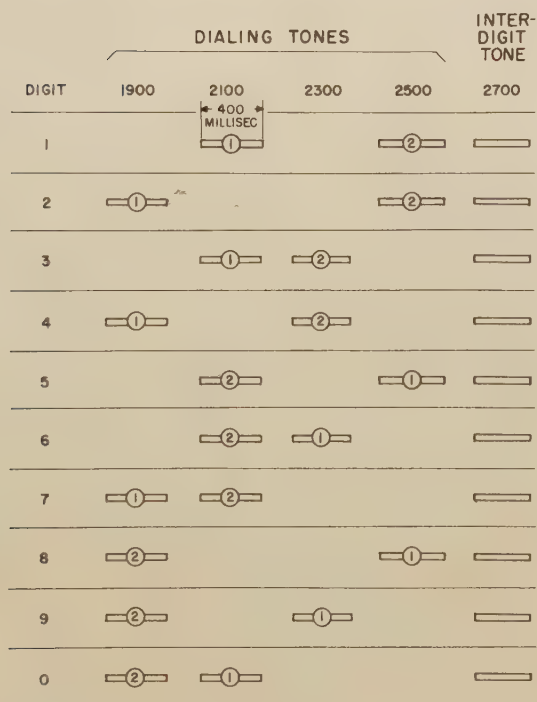


Fig. 4. Sequential arrangement of 2-out-of-4 dialing tones.

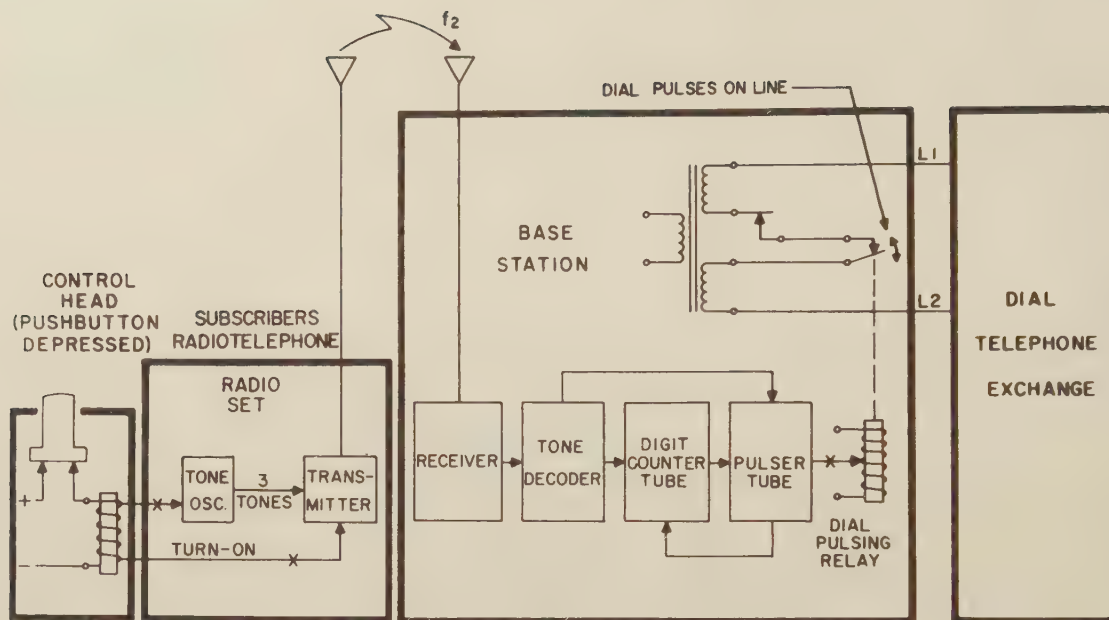


Fig. 5. Pushbutton "dialing" from mobile to exchange.

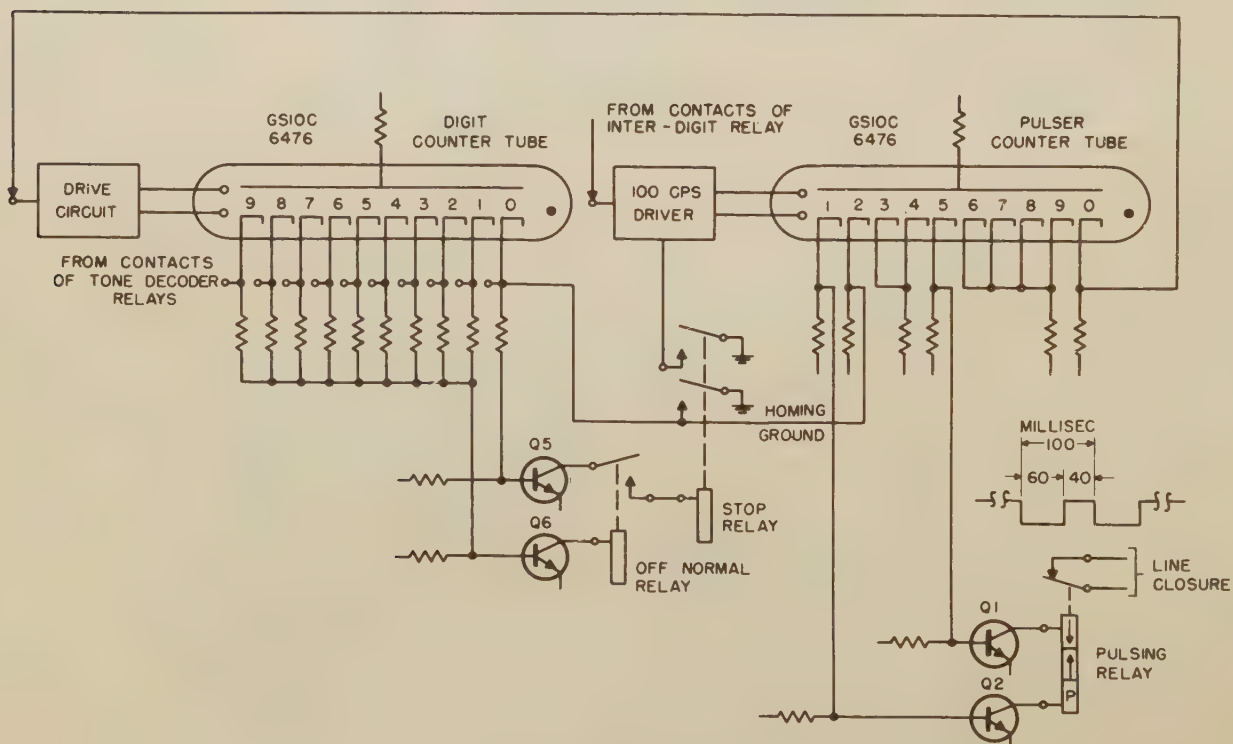


Fig. 6. Digit and pulser counter tubes in control group.

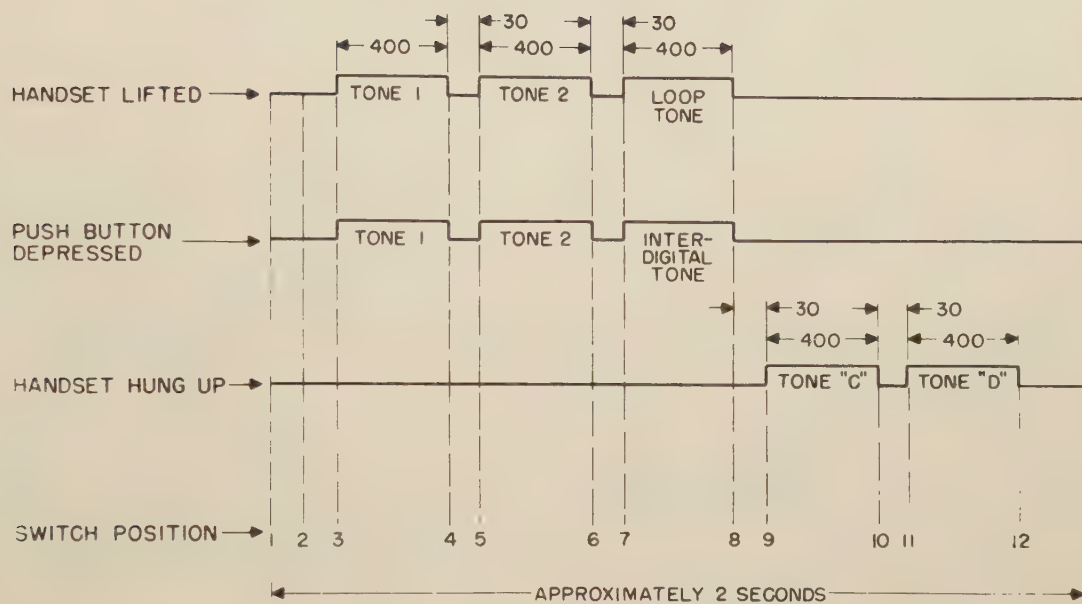


Fig. 7. Supervisory tones from mobile to exchange.

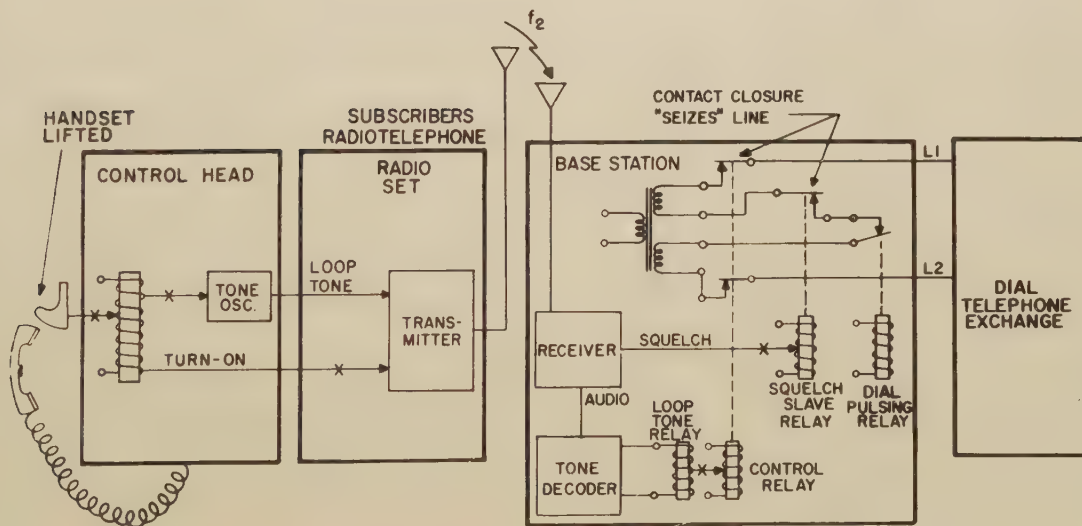


Fig. 8. Mobile initiating call to exchange.

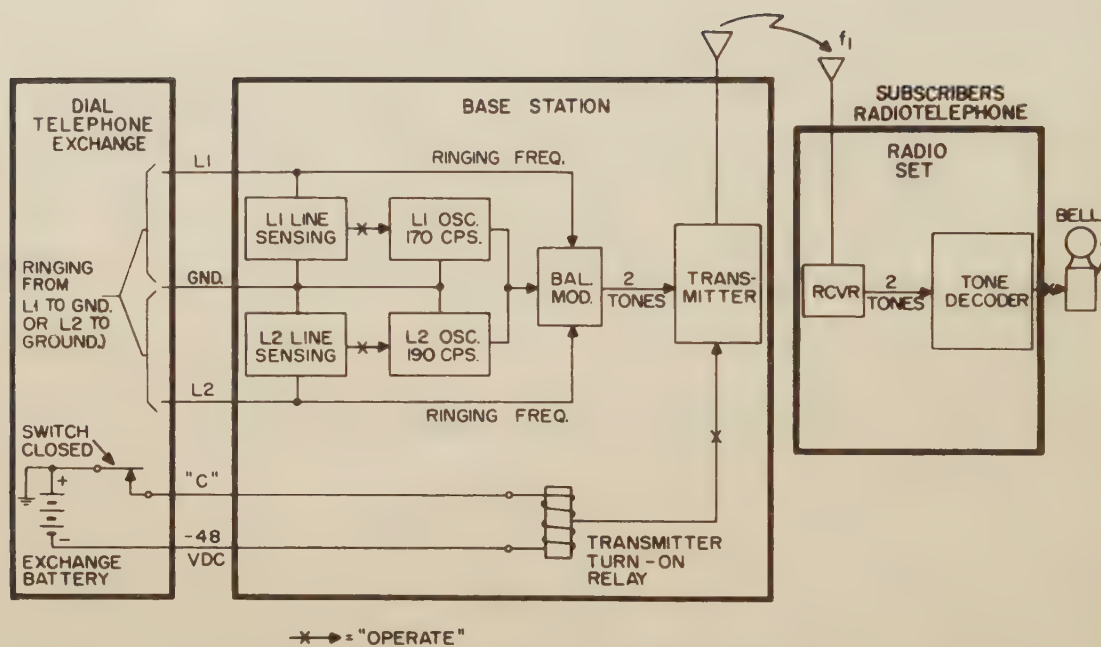


Fig. 9. Subscriber line ringing for selectively calling mobiles.

GUARDED TONE SIGNALLING

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Summary

A method of audio frequency tone signalling which utilizes the maximum number of combinations of a given number of tone channels is described. The principle feature of this method is the use of all tones present as a positive space. Several typical applications of this method to selective calling are described. These include selective party-line telephone ringing, and aircraft selective call.

Introduction

The performance characteristics of tone signalling equipment are mutually dependent upon the task which is to be performed, and the transmission medium over which the signals are to be carried. In general, there are certain requirements which, if fulfilled in the design of such equipment, will permit its application to the majority of selective signalling systems.

The first of these considerations is the code alphabet. In nearly every case, a decimal digit code alphabet is to be preferred for supervisory control and selective calling applications, while an alphanumeric code is preferred for data transmission and combined data/control applications. As a direct result of the choice of a digital code alphabet, code messages consisting of more than one character must be transmitted in sequential form.

The same character may be transmitted any number of times in succession within the structure of a code message.

A minimum number of tone channels should be required to provide the necessary characters in the code alphabet, and should be common to all equipments within the system. In the case of a decimal digit code, the minimum number of channels required is four.

A positive space should be employed between the characters forming a code message, in order to identify a code message as a continuous sequence.

Each character in the code alphabet should be represented by a discrete combination of the available tones.

The partial failure of encoding or decoding equipment which might cause loss of the responses in a single tone channel shall not cause false outputs of a lower order than the desired combination, and

in fact, should cause no output whatsoever.

The three preceding requirements lead to the further consideration that the space signal should be the combination of all tones present, and that the characters be formed by dropping the undesired tones from the space signal. In this manner, the continuity of each code message is preserved, and the failure of a single tone generator or detector will result in lack of response to the space signal; hence no output will result at the decoder.

Individual tone channel bandwidth should be sufficiently broad to accept a spectrum shift as might be experienced in certain transmission media.

The transmission speed in characters per second should be as nearly equal to the maximum keying speed (bit rate) of an individual tone channel, and should take into account a varying amount of phase delay in the different tone channels.

The system should be designed for operation on a time shared basis with voice or other communications, and should not produce false outputs as a result of voice frequency interference.

Encoding equipments should not be limited in the length of a code message that may be transmitted, and decoding equipments should be designed to provide for both serial address, and parallel function decoding in any combination.

The Guarded Tone Signalling equipment described herein has been designed with all of the above considerations in mind, and as a result, is readily adaptable to nearly every conceivable type of control, selective calling, data transmission, or combined system. Due to the broad scope of applications, we must necessarily limit discussion to a relatively few applications; in this case, to the particular application of selective calling.

The principles involved in Guarded Tone Signalling are relatively simple, and focus on the use of the positive space, consisting of all tones used within the system. The use of the positive space in this manner yields a number of advantages over other methods of signalling. Since the presence of all tones is required to form the space, a constant check on the successful functioning of all tone channels used within the system is carried out during the transmission of each code message. The

space signal is used to initiate sampling of the tone channels in the decoder, thus, if all tone channels are not in proper working order, no space signal is derived, hence sampling does not take place.

For the purpose of discussion we shall consider the use of a decimal digit code alphabet. As stated previously, four tone channels are required. There are sixteen possible combinations of four tones. The combination of the four, taken none at a time, will be rejected for possible use as a character, since it may result from failure of the transmission medium, or catastrophic failure of the encoder, or decoder. The combination of all tones will be used as the positive space. Thus, out of the sixteen possible combinations, fourteen are left for use as discrete characters in the code alphabet. Table I shows the code alphabet in terms of these combinations. It should be noted that the letter designated codes will be referred to as "excess codes", since they are not normally used as part of a station address.

While the particular code alphabet shown in the table is based upon the use of four tone channels, the general relationship between the available number of characters in the code alphabet with respect to the number of tone channels required is given in (1).

$$N = 2^n - 2, \text{ where } N = \text{number of characters} \quad (1) \\ n = \text{number of channels}$$

TABLE I

CHARACTER	CODE
0	BCD
1	CD
2	A CD
3	A D
4	AB D
5	B D
6	D
7	B
8	AB
9	ABC
A	BC
B	C
C	A C
D	A
SPACE	ABCD

Encoding

Figure 1 is a time sequence diagram showing the state of the four tone channels during the transmission of the code "1-1-5-A". The solid line represents the presence of the tone. The encoder required to generate such a code is shown in its simplest form in Figure 2. The operation of the manual encoder is simply this: the "on-off" switch is normally open, and as a result, no tones are fed to the line. When placed in the "on" position, all four tones are simultaneously placed on the line, thus generating the space signal. As each of the momentary-acting code switches is operated, the undesired tones are removed, thereby leaving that combination which corresponds to the desired character. Releasing the code switch causes the space signal to be restored. The transmission of a complete code message is accomplished by operating the correct code switches in proper sequence, and returning the "on-off" switch to the "off" position upon completion of the message.

Decoding

The decoding process is primarily dependent upon the transition from the space signal to a code. More specifically, the end of the space causes a sampling process to take place at a predetermined time after the cessation of the space signal, to determine the combination which remains. To better understand the process, we shall consider the block diagram of a typical decoder, as shown in Figure 3. This decoder employs a three character code, the first two digits being serially decoded, thus forming the address, and the third character being parallel decoded, performing a function which corresponds to the character received.

Each of the four tone detectors has both a "zero" and a "one" output in order to simplify the design of the logic circuits. The receipt of all four tones is detected by the space "and" circuit. The space "and" circuit will have an output only when all four tones are present and the lock-out flip-flop is in the reset condition. This output is differentiated and clamped, thus providing a pulse corresponding in time to the end of the space signal.

This pulse triggers a single-shot multivibrator which in turn provides two separate output pulses. The first of these occurs at the instant the single-shot is fired, and is used to set the lock-up flip-flop, thereby preventing any further outputs from the space "and" circuit. The second output pulse is approximately one millisecond in duration, and is delayed from the end of the space signal by approximately ten milliseconds. The delayed output pulse is used as a sampling pulse in the decoding process, and is fed simultaneously to all code "and" circuits. The first code "and" circuit will provide an output only if the combination of tones present at the time of arrival of the samp-

ling pulse agrees with the combination of inputs to which that "and" circuit has been preset. Thus, if the code combination is that to which the "and" circuit is set to respond, an output pulse from the code "and" circuit will cause the first character to register by setting the first digit flip-flop, which will provide a reset pulse to the lock-out flip-flop. Resetting of the lock-out permits the space "and" circuit to start the sampling process at the end of the next space. It is interesting to note that were the received combination not that to which the code "and" circuit is set, registration, and the consequent resetting of the lock-out would not have taken place. The decoder would then have remained locked-out during the remainder of the code message.

Response to the second digit follows the same process, except that an additional input is required in the code "and" circuit, such that the first digit register must be set, in order to register the second digit.

Response to the third digit in the code sequence is identical with the second. Note, however, that four code "and" circuits are parallel such that any one of the four may produce an output, which will correspond to the combination received in the third place.

Selection by Rejection

The decoding process is thus a system of selection

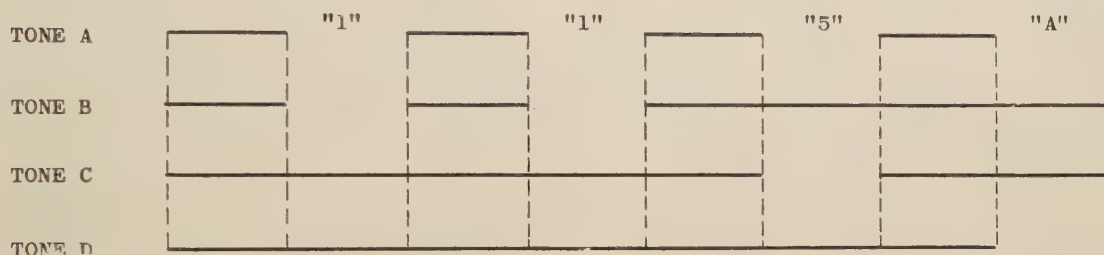


Fig. 1--Time sequence diagram of the code message "1-1-5-A".

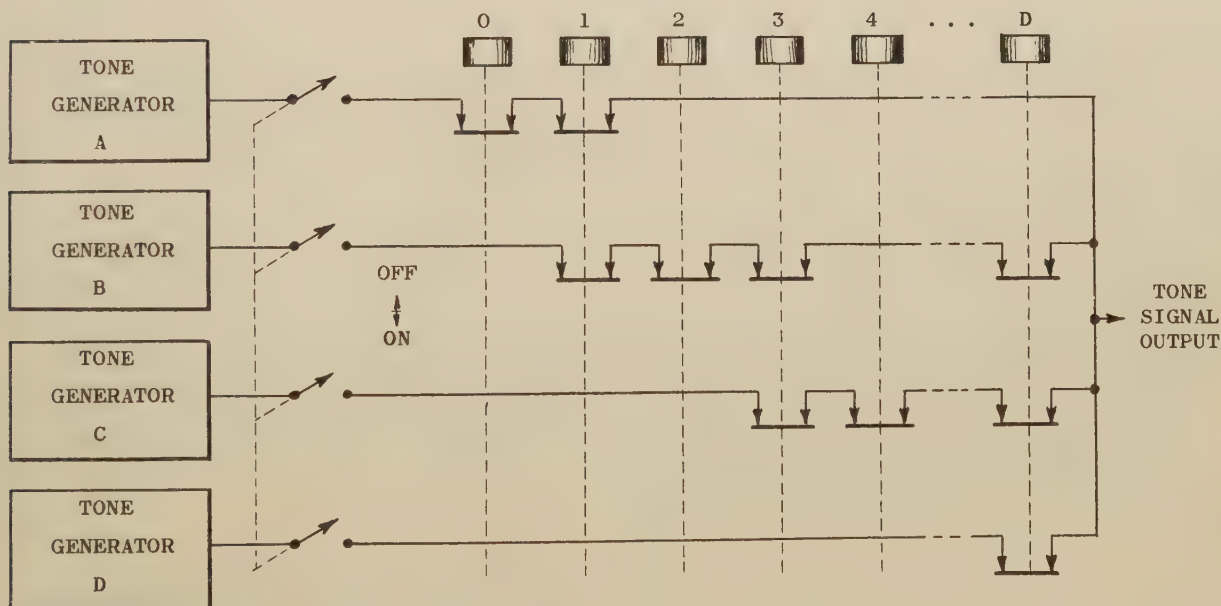


Fig. 2--Block diagram of simplified manual encoder.

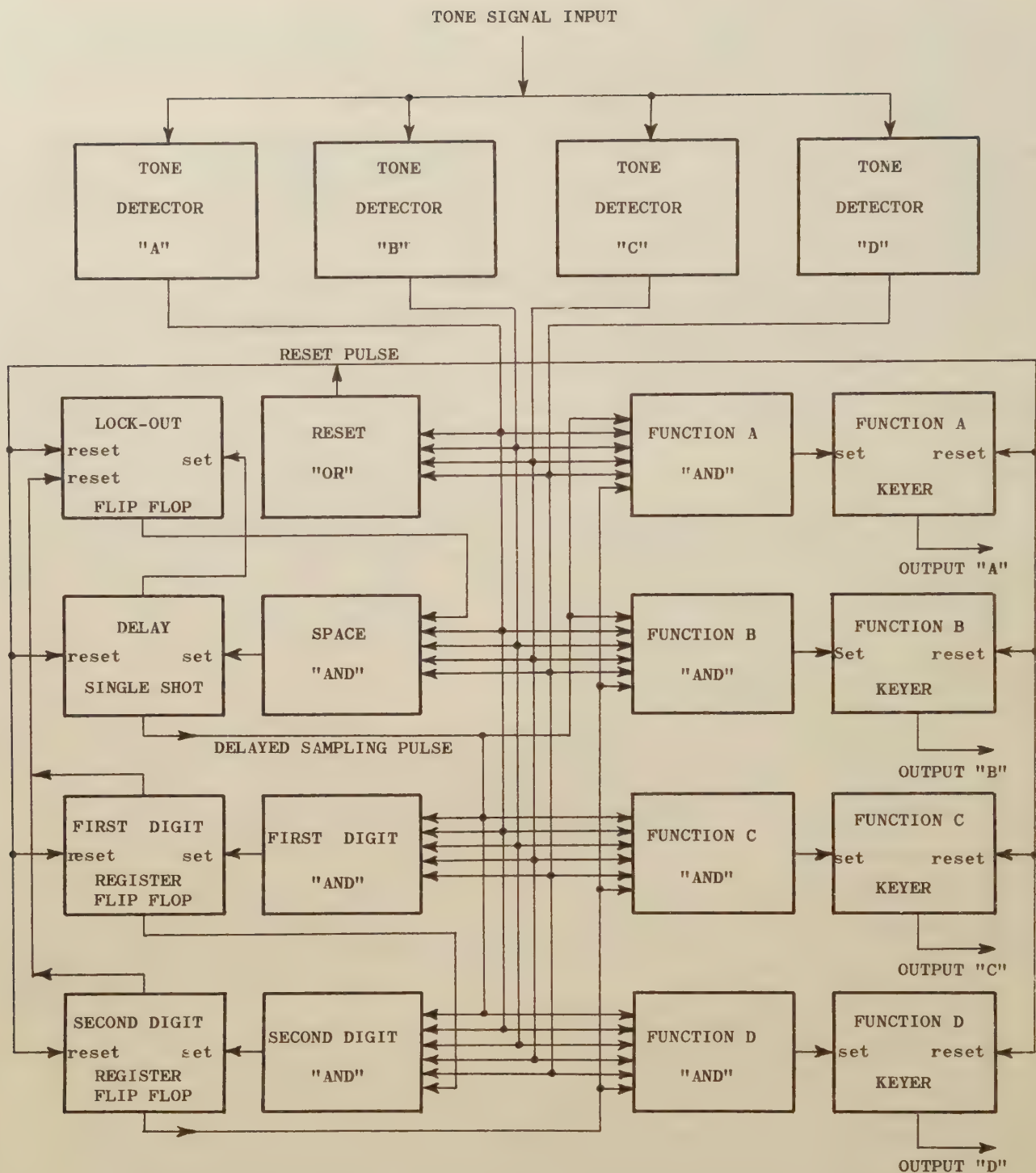


Fig. 3--Block diagram of typical 100 station, 4 function guarded tone signalling decoder.

Selection by Rejection Cont'd.

by rejection. As long as a decoder is in agreement with the code sequence, it will continue to sample the next character. The instant disagreement is reached, the decoder will lock, and remain locked for the duration of the code message. Since each code sequence represents a continuous action in the four tone channels, a reset "or" circuit has to perform the reset function. This occurs at the beginning of each code message.

It should be of interest to point out that in the entire encoding-decoding process, there exists only one critical timing requirement, which is adequately and reliably met with the use of the single shot multivibrator. The duration of the space interval or code interval is of little importance, except that the duration of the code interval must be in excess of the delay time which is inherent in the sampling process. In practice, the operating time of a manual keyboard system such as has been described, is limited only by the operator reaction time in releasing one code switch and actuating another.

The straightforward application of the decoder just described is best illustrated by its actual use in a field test system which is presently installed on the Chesapeake & Ohio Railway Rivanna message circuit. The decoder is used to provide four functional outputs which are: 1) to selectively ring the waystation telephone, 2) to cut-through the Rivanna message circuit to the Clifton Forge message circuit, 3) to apply 20 cycle ringing on the cut-through circuit, and finally, 4) to disconnect the cut-through.

As proof of the unique adaptability of the Guarded Tone Signalling method, consider this example. As long as the last characters are restricted to excess codes, that is, combinations which are never used in any address, then the three digit decoder will never respond to a code message either longer or shorter than the prescribed three digits, even though a longer code message may contain the entire three digit code within itself. To illustrate, let us assume the decoder is set to respond to the code "1-1-A". Transmission of the code "1-1-1-A" will result in the decoder locking-out upon receipt of the third digit "1" in the address. Thus in a system such as described, we may successfully add three digit decoders in an existing two digit system, without having to retrofit the existing decoders.

Group Call

As a second example of the versatility afforded by this signalling method, let us consider the requirement for group call, and how it can be provided. Referring once again to the decoder block diagram of Figure 3, it can be seen that the four function outputs are prevented from operating un-

less the second digit register has been set immediately prior to the receipt of the function code. You will further note, that each time the decoder agrees with the received code character, one more sampling pulse will be generated. In fact, all decoders generate one sampling pulse corresponding to the first character in every code message. Those that agree with the first character generate a second sampling pulse, and so on. If we merely remove the requirement that the second digit register be set in order to actuate the function codes, and further stipulate that function code characters be excess codes, that is, not used in any address, then we have enabled the decoder to accept the function code at any point in the code message. If we then transmit a single character code message consisting of the function code only, all decoders will perform that function, since all decoders generate the sampling code on receipt of the first character in any code message. By transmitting a two character code message, consisting of an address digit followed by the function code character we will perform the function code only at those decoders which agreed with the first digit, since all decoders set to respond to a first address digit other than that which was transmitted will have locked-out on the first part of the code message, and consequently will not generate a sampling pulse for the second character in the message. Thus, we have provided a decade group call. In a five digit address system of selective calling, there is one group of 100,000, 10 groups of 10,000, 100 groups of 1,000, 1,000 groups of 100, 10,000 groups of 10, and finally, 100,000 unit calls. It should be pointed out that, as in the preceding case, no decoder is capable of responding to a code message which has an excess of address digits. It is, however, necessary that all decoders used within the system contain the same number of digits in the address, except in a special case. That case is best illustrated by a typical volunteer fire system, equipped with selective call. Assume that 21 volunteer companies share the same county fire radio system. By using equipment such as described we may assign two digit decoders to each of the volunteer fire houses for the purpose of sounding the siren, and three digit decoders to each of the volunteer firemen. Decoders supplied to the individual fire company members will bear the same first two digits in their call as the fire house decoder. By transmitting a two digit address, the siren will be sounded and all members of the company will be group called. By transmitting a three digit address, each individual company member may be selectively called without sounding the siren.

Rapid Code Change

In those cases where it may be necessary to change the address to which the decoder will respond, this may be done with ease. Since the response of the

decoder is determined by "and" circuits, we merely have to change the inputs to correspond to the desired response code. This may be accomplished by the addition of a rotary selector switch associated with each of the address "and" circuits. Airline selective calling systems have particular need for this feature.

Failure Alarm

In addition to the need for changing the codes rapidly, an airline selective calling system must also feature a failure alarm. Again, due to the unique capabilities of Guarded Tone Signalling, this may be accomplished with relative ease. The purpose of the failure alarm is to indicate to the pilot that his selective call system is in failure, and as a result he must aurally monitor his radio. Such a failure alarm is practical because all calling code messages are based on a common language, and more especially all code messages within the system utilize the common space. All code messages are also detectable since they are represented by a continuous activity in one or more of the system's tone channels. As we have studied, the reset "or" circuit of the decoder in Figure 3 has an output available which is continuously present during the reception of any code message. Thus,

by comparing the output of the "or" circuit with the output of a second uninhibited space "and" circuit, we can derive the failure signal. In other words, if we sense a code message, yet never sense the presence of a space within that code message, then either the ground station encoder, or the aircraft decoder is in partial failure. We may detect this failure at the called station, and alert the pilot. While it is true that catastrophic failure of the encoder or decoder is not provided for in the abovementioned failure detection scheme, we must remember that such failure may be detected by other means, such as, supply voltage alarm, input level alarm, receiver alarm, etc.

Code Readout

Another byproduct yielded by the use of the common language code message is typified in the application of Guarded Tone Signalling in supervisory control applications. In a process control system, where it may be necessary to record the sequence of events as they occur, a special type decoder, which will record each code message transmitted within the system on a standard IBM Output Writer, should be used. The relative ease with which this may be

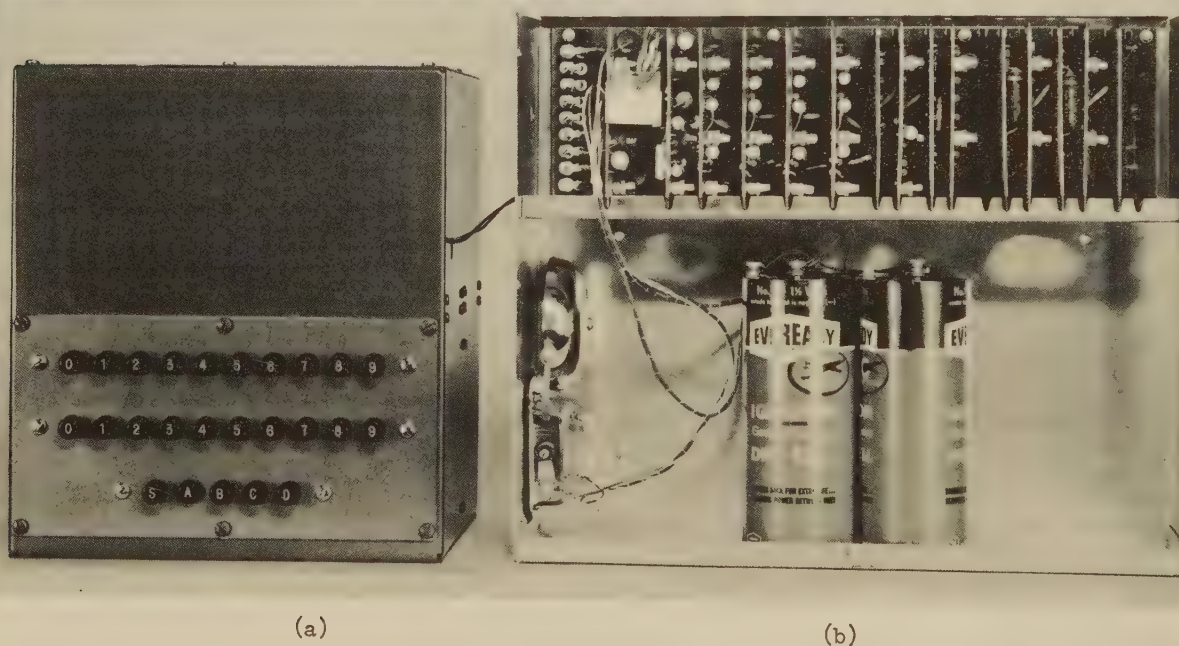


Fig. 4--Typical guarded tone signalling equipment. (a) 100 station, 4 function semi-automatic encoder, battery-powered. (b) 100 station, 1 function battery-powered decoder.

done is again attributable to the use of the positive space, the continuity of individual code messages, and the common code alphabet of the system. Incidentally, you will observe that the digital alphabet shown in Table 1 is arranged in a minimum error code, thus taking into consideration its use in systems where normal analog to digital conversion techniques are employed.

The foregoing examples have illustrated the versatility of the Guarded Tone Signalling method. As indicated previously, field test systems are presently in everyday use. The theoretical concepts have been reduced to practice, and another useful tool will shortly be available to the system

designer in the fields of supervisory control, selective calling, and medium speed data transmission.

Typical equipment is illustrated in Figure 4. The encoder and decoder shown, are designed to operate in a 100 station party line ringing system at a nominal speed of 10 digits per second. They may be interconnected by any voice circuit, including single sideband suppressed carrier.

Continuing development of equipment, for applications other than mentioned here, is in progress... ..and the end of this program is nowhere in sight.

15 KC SPLITS, - A SOURCE OF CHANNELS IN THE 150 MC BAND

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The major problem facing the mobile radio industry continues to be that of finding enough radio frequency channels to accommodate all the applicants who need and desire mobile radio facilities. User groups and industry have kept the FCC well informed on our needs for additional spectrum space, and we believe they have earnestly tried to fulfill them, but it will always be incumbent upon us to make the maximum possible use of the spectrum space we have. Historically, the Land-Mobile Services have made very judicious use of this important national resource. Channel splitting is a fine example of more effective utilization of the spectrum. The next move in this direction is undoubtedly the implementation of the 15 KC splits, - the so-called tertiary frequencies in the high-band.

We are suffering real channel congestion right now. In metropolitan areas the business radio channels, for example, are loaded about 10 systems deep and this explains why tone squelch systems are finding increasing application. Tone squelch systems are a panacea but they cannot replace the communication effectiveness of a clear radio channel. Tone devices simply prevent you from hearing other systems on the frequency, they do not provide air time when the other fellow is using the channel nor do they prevent capture of the frequency by a powerful signal.

With so many systems on any one frequency, the channel congestion becomes almost intolerable. In addition, it has the serious effect of reducing the Mobile-to-station range to as little as a few miles. This occurs because the Mobile units of one system have to compete with the powerful base transmitters of other systems in order to be heard. This brings up the point that not only should we consider the early implementation of the tertiary frequencies, but we should also consider the more wide-spread use of two-frequency Mobile systems, wherein a portion of the available frequencies are reserved for Mobile use only so that Mobiles will not have to compete with base transmitters. The taxi-cab industry is a classic example of the high traffic handling capacity of two-frequency systems under very congested metropolitan conditions. Without two-frequency operation, it would not have been possible for taxi-cab radio to do the job it did in the days when only four channels were available to any area, - large city or small. The success of taxi radio operations points up the desirability of considering reservation of some of the tertiary frequencies for Mobile use only. This is a means by which some of the present interference in metropolitan areas could be alleviated.

The specific questions we would like to consider in this paper are these: Just how satisfactory are the tertiary frequencies? Is it better to choose a tertiary frequency or to share a channel with one or more systems? We have considered these questions from both a laboratory and a practical operation standpoint, and are pleased to report on our findings.

One of our affiliated Sales and Service Agencies Communications Engineering Company of Dallas, Texas, recently installed what we believe to be the first industrial system placed in operation on a tertiary frequency, the Go-Crete Ready-Mix Concrete Company of Dallas and Arlington, Texas. The Go-Crete system has been in and working since November, 1959 and engineers of both CECO and General Electric have conducted field surveys, independently, to determine the degree of interference which would occur between that and systems 15KC removed on each side of its operating frequency. We might add at this point that the Go-Crete system uses no form of tone protection.

Before discussing the Go-Crete tests and the results of those test, we would like to show a series of curves which will show the engineering considerations of operation on 15 KC splits.

Figure 1 shows the spectrum of energy radiated by a transmitter which meets the FCC narrow band standards, under three conditions of modulation; that is, no modulation, average modulation, (which is voice modulation just below the limiting level), and heavy modulation, (which is voice modulation sufficiently heavy to result in 10 db of limiting.) These curves show that, when a transmitter is voice modulated plus and minus 5 KC, the radiated energy will be fairly flat out to the plus and minus 5 KC points and then trail off at about a 6 db per KC rate. The slope of the transmitter side-band envelope determines the slope of the receiver selectivity slope of more than 10 db per KC would not provide improved interference rejection.

It will be seen that the transmitter side-band envelope is the limiting factor in channel spacings of 15 KC. If we assume receivers with a pass-band of 10 KC, centered at the 15 KC points on each side of zero frequency, we observe that a considerable amount of energy falls into the pass-band of those receivers from such an interfering transmitter. Therefore, the only way to avoid interference is to provide sufficient geographic spacing so that the unwanted energy is sufficiently attenuated. We cannot expect a receiver to reject energy that falls directly in its pass-band.

When the transmitter is unmodulated, the carrier is a spike whose maximum amplitude is zero db as shown on the ordinate of the graph. The noise spectrum starts at 46 db down and trails off to about 80 db at the 15 KC points. In 30 KC channel spacing, the transmitter modulation envelope is not the limiting item in adjacent channel operation. The limiting item then is the noise spectrum of the offending transmitter.

Figure two shows the true selectivity of two types of receivers that were used in the Go-Crete tests. The 100 db specification usually associated with the so-called 20 db quieting method of measuring selectivity will not be found on these curves. The 20 db quieting method of measuring selectivity is fallacious because it ignores the desensitizing effect of a strong off-channel signal. Only a single signal is used in that method. These curves were plotted on the basis of 6 db degradation of a 12 db sinad ratio resulting from the application of an unmodulated interfering signal. The sensitivity of the receivers is approximately half a microvolt for 12 db sinad at the output. The curves show that a quarter of a microvolt of undesired carrier at zero frequency, (the frequency to which the receiver is tuned), will degrade the 12 db sinad, 6 db. As we move off resonance 15 KC we see that approximately 50 microvolts of undesired carrier is required to degrade receiver B to that degree and more than 5,000 microvolts for receiver A. At 30 KC off resonance - the adjacent channel - both of these receivers exhibit greater than 90 db rejection.

The standard EIA two-signal method of measuring selectivity uses modulated signals for both the desired and undesired signals. However, in these curves no modulation was used on the interfering signal generator in order to plot a selectivity curve that could be used as a pattern against the transmitter side-band envelope to determine the points at which interference would occur. As mentioned above, these curves, at the 15 KC points in which we are particularly interested, show 42 db rejection for receiver B and 82 db rejection for receiver A. Standard EIA curves would show receiver B to have 24 db attenuation and receiver A, 48 db attenuation. In other word, the effect of using a modulated signal for the interference would be to displace the curves about 3.3 KC due to the modulation.

Note that the slope of the selectivity curve of receiver B is approximately five db per KC while of receiver A it is ten db per KC. Our examination of a transmitter side-band envelope indicated that it would be desirable for a receiver to have a steeper slope than that of receiver B, and this proved to be the case in the Go-Crete Tests, although receiver B gave a good account of itself under these very adverse conditions.

Figure three shows the interference situation by combining the selectivity curve of receiver A with the transmitter spectrum under conditions of average modulation and no modulation, when transmitters and receivers are exactly on frequency.

A second set of transmitter curves shows the conditions prevailing when they are moved one and one half KC closer to the receiver frequency. This would be the condition when the transmitter and receiver are at the limit of 0.0005% tolerance.

Interference will occur where the transmitter and receiver curves intercept. Looking first at the transmitter curves which depict exact frequency operation, it will be seen that the noise spectrum of the unmodulated transmitter first starts to enter the receiver pass-band when the interfering transmitter is 72 db stronger than it would need to be to cause interference on the frequency to which the receiver is tuned. In other words, by offsetting the interfering transmitter 15 KC, we obtain 72 db increased protection against interference. With modulation on the interfering signal however, we have a situation shown at left where the protection is only 36 db greater than co-channel. As mentioned earlier it will be seen that the voice modulation starts to intercept the receiver pass-band at the 36 db point. The only way we could improve this situation would be to reduce the noise width or pass-band of the receiver and this in turn would require that we modulate the desired transmitter correspondingly less than plus and minus 5 KC.

Looking at the transmitter curves which are only $13\frac{1}{2}$ KC displaced from the receiver, we note that the noise spectrum is not now the limiting item but rather, the carrier spike which starts to enter the receiver pass-band at a point 67 db above the zero level. Similarly, we have less attenuation of the modulated undesired transmitter. Our rejection has fallen from 36 db to 27 db. This points up the increased importance of maintaining exact operating frequency of transmitters and receivers in 15 KC split-channel operation.

The interference relationships shown by this set of curves were born out in the Go-Crete tests. It was found that the interfering carrier produced no interference until our test vehicle was very close to the offending transmitter. However, one to three miles of separation was required to avoid interference from voice modulation peaks. As the test car approached the offending transmitter, voice peaks could cause squelch bursts in/or degrade receiver sensitivity to a desired signal. The degree of interference shown by these curves varies all the way from 27 db to 72 db, - a difference of 45 db. Since radio propagation over plain earth drops off at a rate of approximately twelve db per octave, this would indicate about a ten to one difference in the range in which interference could occur when the transmitter is modulated, as compared to when it is not modulated. If one encountered carrier noise interference at 0.4 of a mile he could expect detectable interference from voice peaks out to four miles. Fortunately, the voice peak type of interference is not destructive nor is it especially annoying. It could be eliminated by the use of tone operated squelch.

Figure four shows the interference relationships prevailing with receiver B. As might be anticipat-

ed, this receiver, with a skirt selectivity slope of only five db per KC does not provide the rejection of a receiver having a ten db per KC slope. In this case the offending carrier produces considerably more interference than in the case of receiver A. It is down only 41 db at 15 KC and 34 db at 13½ KC. However, the receiver does a good job with respect to the modulated interfering signal, where receiver A has an attenuation of 36 db, this receiver has 30 db. At the 13½ KC points where receiver A provided 27 db attenuation, receiver B provides 24 db, these relative relationships were born out in practice. Receiver B did show greater desensitization due to an interfering carrier than receiver A, but it still did a good job of rejecting the unwanted modulated signal.

Figure five shows the severe loss of mobile talk-back range when a desired and an undesired station are operated on the same frequency (co-channel) within five to fifteen miles of each other as compared to a situation wherein the two stations are given a separation of 15 KC. Mobile units in each case operate on the same frequency as their associated base station. Without an interfering station on the air the normal range over which a 50-watt station will deliver a one microvolt signal to the mobile receiver (or vice-versa) is 23 miles.

The lower curve shows that the loss of talk-back range due to the operation of a co-channel interfering station is quite severe. If the stations are five miles apart, the talk back range to the desired station is only 1.6 miles. With 15 miles spacing between the two stations, it is only 3.2 miles.

When the two systems are spaced by 15 KC however, the mobile talk-back range under average modulation conditions is nearly as great as though there were no interfering station on the air at all. Even when the two stations are only 5 miles apart, the talk-back range is reduced only from 23 to 13 miles under average modulation conditions. Summarizing, it can be seen that there is about an eight to one improvement in talk-back range using 15 KC spacing as compared to co-channel operation. There is at least a five to one improvement even when the interfering station is modulated well into the limiting region. If one had a choice of operating on the same channel with another system, or of selecting a tertiary frequency, it can be seen that the obvious choice would be in favor of the tertiary frequency.

Figure six shows the range over which a mobile unit would receive interference as it approached a 15 KC adjacent-channel station under conditions of no modulation, average modulation and heavy modulation of the interfering base transmitter.

If the desired and undesired base stations were separated by 15 miles, the undesired carrier would cause interference over a radius of half a

mile, while average modulation would interfere over a radius of 2.6 miles. Heavy modulation would be detectable somewhat beyond 3½ miles from the interfering base transmitter.

Note that the closer the two stations are located to each other, the smaller the range over which interference to mobiles will occur. This is obviously true of separation prevails. On the other hand, if the base stations are too close together, they will interfere with each other. In the Go-Crete - Uvalde systems, base stations are 9.3 miles apart and at that distance, do not interfere with each other.

Figure 7 is another way of looking at the comparison between co-channel versus 15 KC and 30 KC offset operation. We see that two co-channel 50 watt base stations with 100 foot towers will interfere with each other at a distance of approximately 56 miles. If they are displaced 15 KC from each other, the interference range drops to 15 miles, while with 30 KC spacing it is less than 0.2 of a mile. Base-to-mobile interference range, co-channel, starts at 29 miles, while at 15 KC displacement, interference drops to 3.5 miles. With 30 KC displacement interference range is less than 0.2 of a mile.

The tabulation also includes low-band comparisons based on co-channel operation and 10 KC and 20 KC splits.

Figure 8 is a map of the Dallas-Fort Worth area showing four stations in the special industrial radio service operating within 15 KC of each other. The Go-Crete station around which we made the field tests is in the lower right hand portion of the map and operates on 152.885 megacycles, a tertiary frequency licensed under a developmental grant. It employs a 50-watt transmitter and has a 100 foot steel tower, but some of the elevation is not realized because the plant is located in a natural depression. A 5.8 db gain omni-directional antenna is employed with approximately 150 feet of RG 8 U transmission line. We credit this station with an effective radiated power of 50-watt for purposes of evaluating our tests. Another Go-Crete station in the system is located approximately 20 miles to the west at Arlington. Northwest of the Go-Crete station at Dallas, is the Uvalde base station operating on 152.900 mc, that is only 15 KC removed from Go-Crete. It has a 150 foot tower with a 5.8 db omni antenna and enjoys somewhat higher ground elevation than the Go-Crete base station. Also, RG 17 U transmission line is employed.

Approximately 20 miles to the west of the Go-Crete - Arlington station, Consolidated Vultee has a base station on a 152.870 mc, which is 15 KC below the Go-Crete frequency.

After the Go-Crete installations were completed still another system was discovered to be operating in the area on the frequency 152.87 mc at Garland, Texas which is about 15 miles northeast of the Go-

Crete, Dallas transmitter. It is understood that this system, owned by the Cooper Concrete Company, is still using a wide-band base station receiver and the receiver does respond to the Go-Crete transmissions as might be expected. However, the Cooper Company have not complained and a simple solution to the problem will be the installation of a narrow band base receiver.

The stations between which we would expect the most interference are those of Go-Crete Dallas and Uvalde. The actual distance between the two is 9.3 miles and we measured a propagation loss between the two stations of 119 db. Over plane earth for 9.3 miles we would normally expect about 107 db attenuation. The additional loss in this case is probably due to the radio path traversing the heavily built-up down-town area of Dallas.

To the best of our knowledge there has been no complaint of interference between any of these systems in the Dallas area save that of the Cooper wide-band set. Undoubtedly the mobile units can and do get into interference situations as they approach other system base stations but these cases are sporadic and probably go unnoticed.

Three series of field tests were made using the Go-Crete Dallas station as a source of interference. A station wagon was equipped with laboratory test equipment and two complete mobile combinations which provided for transmission and reception of the three frequencies involved in the Dallas area pertinent to this study. One of the test mobile receivers (receiver B) was designed for operation on 30 KC channel spacing, - the other, still more selective, was designed for either 30 or 15 KC channel spacing (receiver A). The actual selectivities of these two receivers are as given in figure 2.

In the first series of tests, the objective was to determine the degree of interference the Go-Crete station would cause mobile receivers operating 15 KC above and below the Go-Crete frequency in the absence of any desired signal with the test car preceeding away from the Go-Crete station.

The 15 KC receiver (A) experienced squelch opening on voice peaks only at 0.4 miles. At one mile this effect had disappeared and receiver A was free of interference.

The 30 KC receiver (B) reproduced understandable voice ou to one mile. Voice peaks caused intermittent squelch opening at two miles.

In the second series of tests, the objective was to determine the degree of interference the Go-Crete station would create in mobile receivers 15 KC above and below its operating frequency while those receivers were responding to a weak desired signal applied by a local signal generator in the station wagon. The level of the desired signal was 1.8 microvolts modulated at 1000 cps for easy recognition.

At 0.4 miles receiver A was measured to have a protection of 60 db against the interfering signal. Receiver B showed 48 db protection. At one mile receiver A provided 48 db protection; receiver B 43 db. At the two mile test points receiver A was free of any measurable interference while receiver B provided 44 db protection and required $4\frac{1}{2}$ microvolts to deliver a clear, thousand cps modulated, desired signal.

At 3 miles receiver B was measured to provide 40 db protection and 3.6 microvolt of desired signal was required to completely override the interference. This concluded the second series of tests. It is estimated that receiver B would be completely free of interference at 3 $\frac{1}{2}$ miles.

In the final series of tests, the objective was to determine the effect of operation beyond the 0.0005% frequency tolerance of present day narrow-band transmitters and receivers. This would add up to 1 $\frac{1}{2}$ KC at 150 MC but we actually returned the receivers 2 KC closer to the interfering frequency. As expected, off-frequency operation rather seriously degrades the interference rejection capabilities of 15 KC systems. The 6 db slope of the transmitter spectrum indicates that the degradation should be about 6 db for each KC off-frequency. Receiver A was free of interference at about 3 $\frac{1}{2}$ miles but receiver B still experienced slight interference at a 4.8 mile test point.

The most serious source of interference in an FM signal is the carrier frequency, since it is the strongest discrete frequency and because it is present throughout each transmission. If the carrier does create interference, that interference is constant for the duration of a transmission. Interference due to voice modulation side-bands, on the other hand, is pulsating in nature and while it may be annoying, it often does not prevent the listener from understanding a desired transmission.

The use of tertiary frequencies is much to be preferred over co-channel operation because the carrier is off-set 15 KC to a point where it falls on the knee of the selectivity curve of a well-designed receiver. At the knee, the well-designed receiver will have an attenuation of approximately 80 db. Although 15 KC displacement is not sufficient to prevent a portion of the modulation side-bands from entering the receiver pass-band, nonetheless, considerable protection is afforded. The curves on figure 3 show that under on-frequency operating conditions, tertiary frequencies will provide 72 db protection as far as the carrier (due to it's noise spectrum) is concerned and 36 db protection against the modulation sidebands. Thus we see that we have the maximum protection to the interfering component which is the most troublesome and the most constant. Where we have the least protection, the modulation side-bands, some additional protection can even be obtained by encouraging out neighbor not to modulate so heavily. Furthermore, the interference we do obtain is pulsating or intermittent in nature and desired messages can still be read in between the inter-

ference peaks.

Inasmuch as the interference level varies all the way from 36 to 72 db, we can expect a similar wide variation in the range over which interference occurs. For example, if the carrier noise created interference over a radius of half a mile from an interfering station, the modulation peaks can cause interference anywhere from $2\frac{1}{2}$ to $3\frac{1}{2}$ miles. And these laboratory predications were very well borne out in the Go-Crete field tests.

Another way of comparing co-channel operation with 15 KC adjacent channel operation is to say that with 15 KC spacing, interference is reduced by 36 to 72 db and this is equivalent to reducing the output of a 50-watt interfering transmitter from 50 watts to 13 milliwatts or from 50 watts to 4 microwatts.

Conclusions

1. The implementation of 15 KC split frequencies in the 150 megacycle band is entirely feasible, practical, and in the interest of good spectrum utilization, provided only that sufficient geographic spacing is employed between base stations. Equipment meeting the present FCC standards for narrow-band operation is entirely adequate for this type of operation.

2. With receivers having a two-signal selectivity of 40 db or more at plus and minus 15 KC, the limiting factor on interference range is the side-band envelope of the interfering transmitter. But even with receivers having only 20 db of selectivity at the 15 KC points, 15 KC split-channel operation is clearly superior to co-channel operation.

3. Tone-operated squelch will be even more effective in reducing interference with 15 KC spacing between the systems than with co-channel operation because the desensitizing effect of an interfering carrier is reduced by as much as 72 db. Tone operated squelch can be expected to completely eliminate nuisance interference due to voice modulation peaks.

The supporting engineering data and performance curves in this paper were supplied by N. H. Shepherd, Advance Engineering Section, Communication Products Department, General Electric Company, whose assistance is gratefully acknowledged.

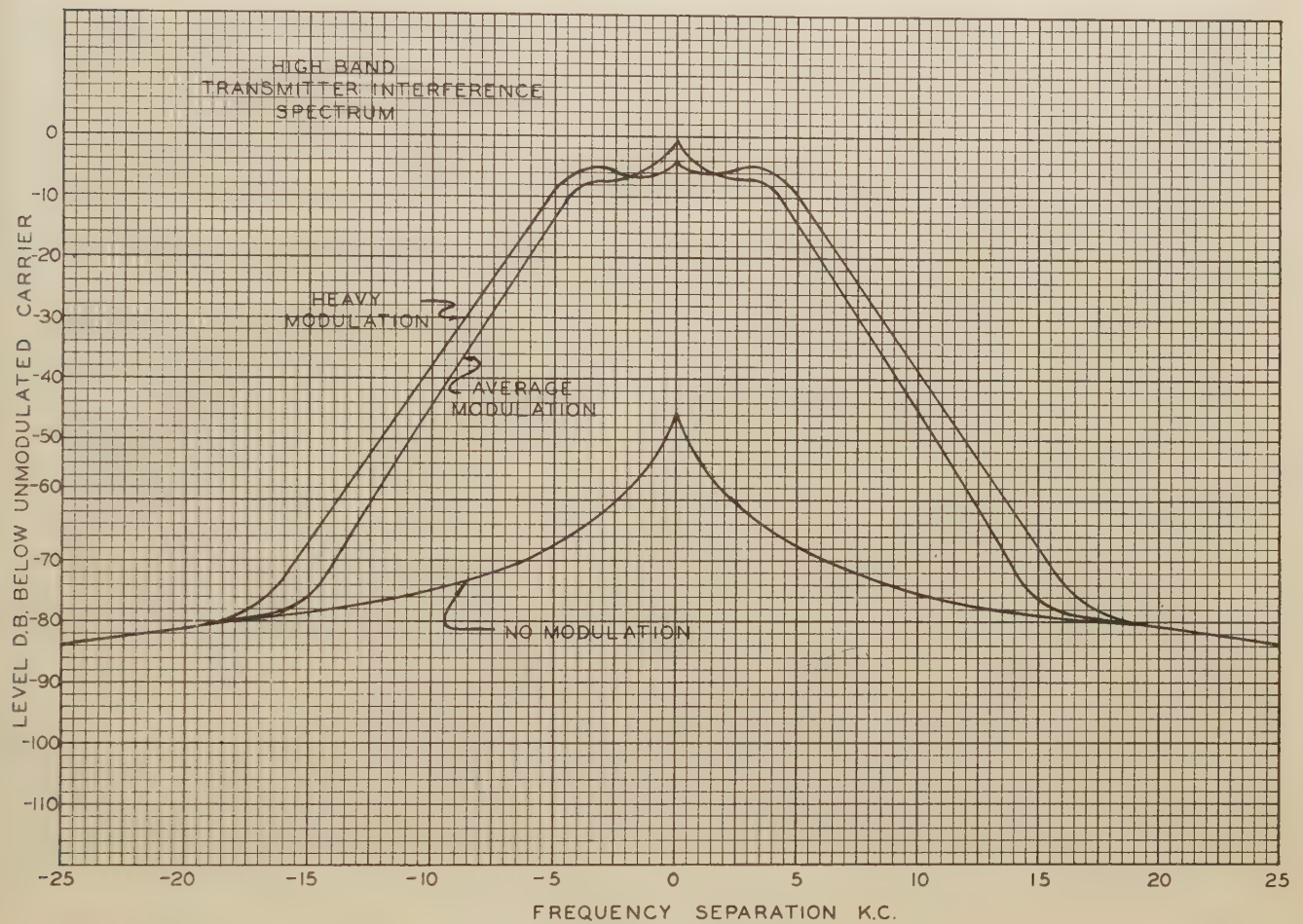


FIGURE 1

High Band Transmitter Interference
Spectrum

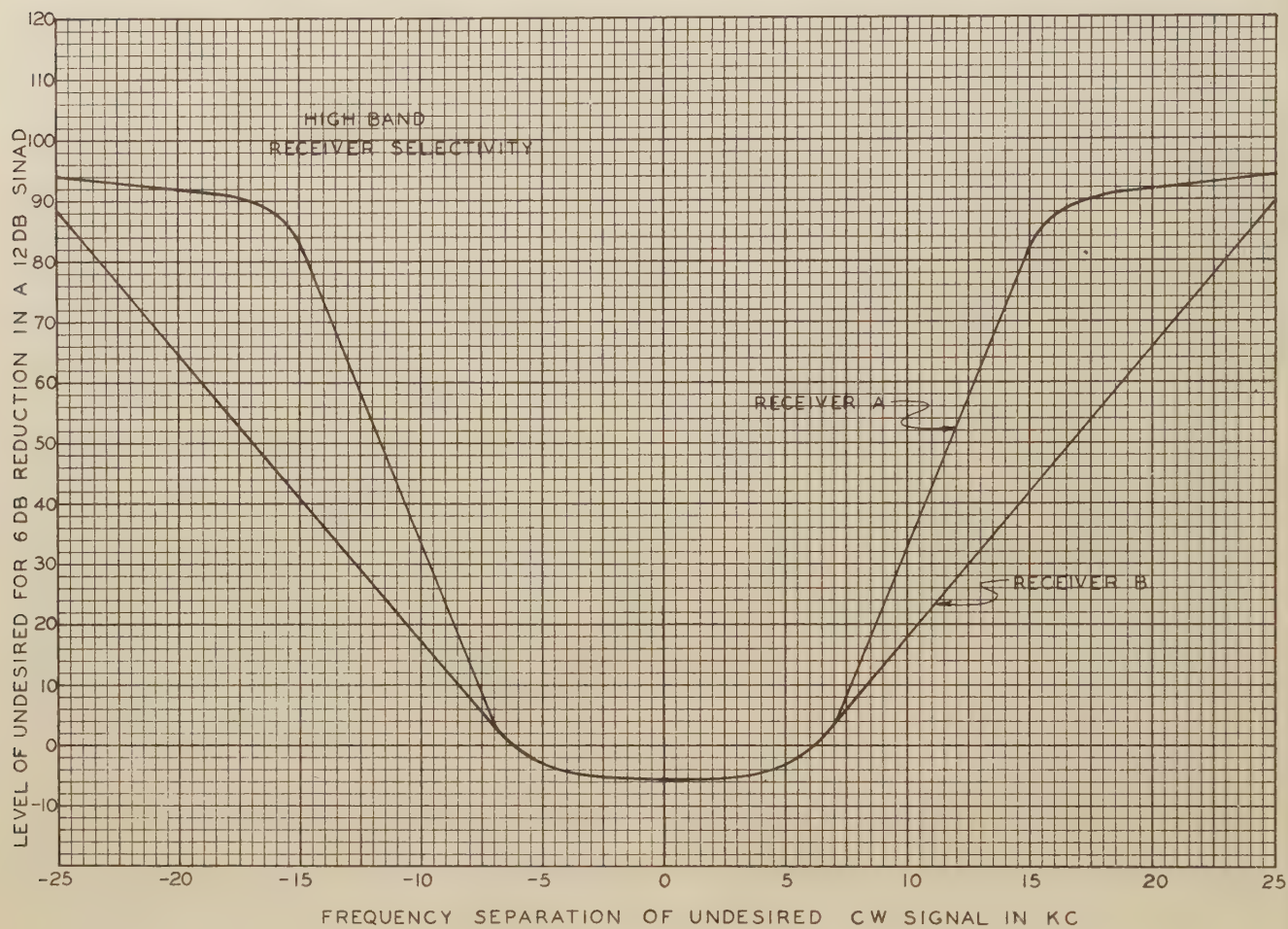


FIGURE 2

High Band Receiver Selectivity

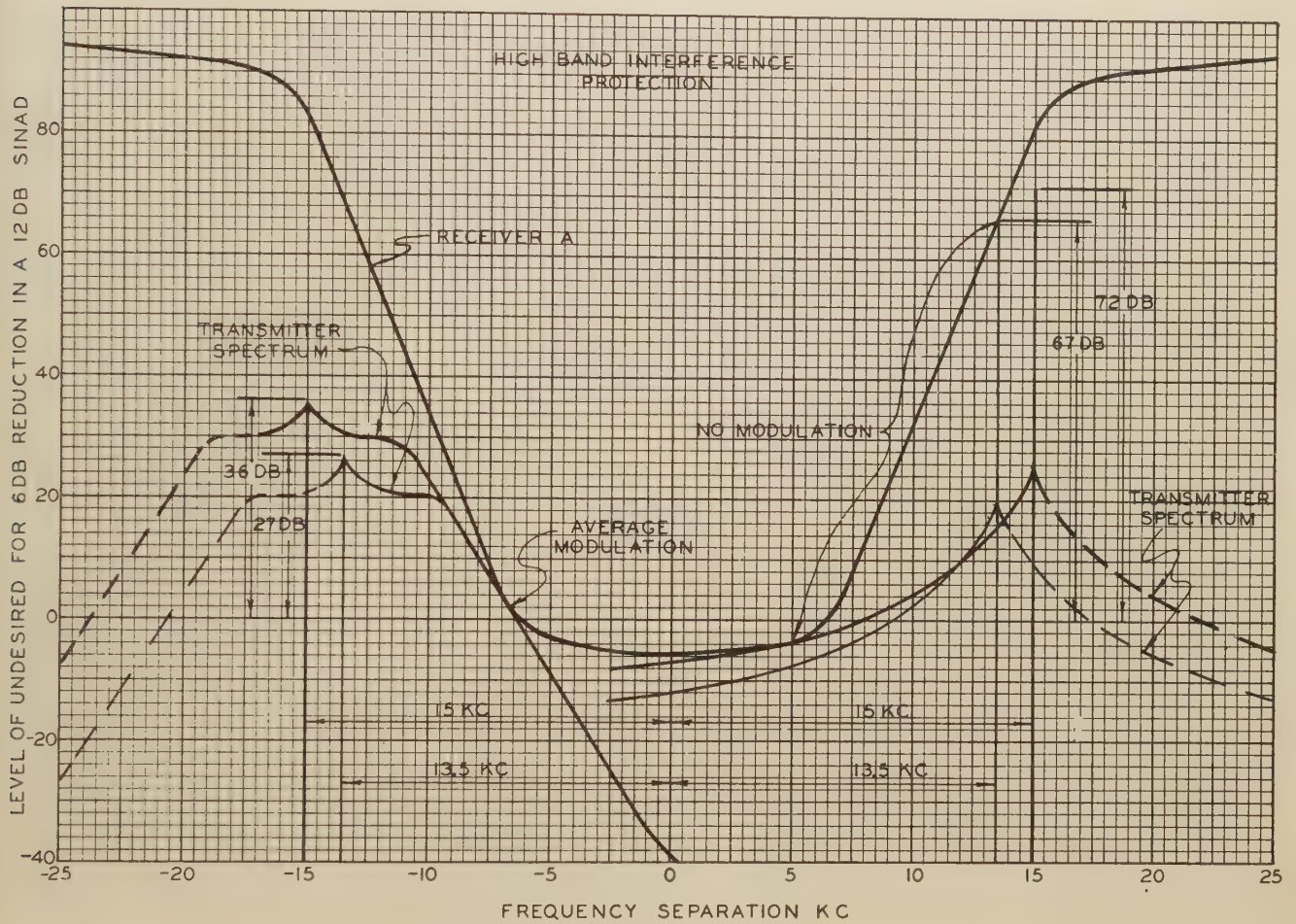


FIGURE 3
High Band Interference Protection

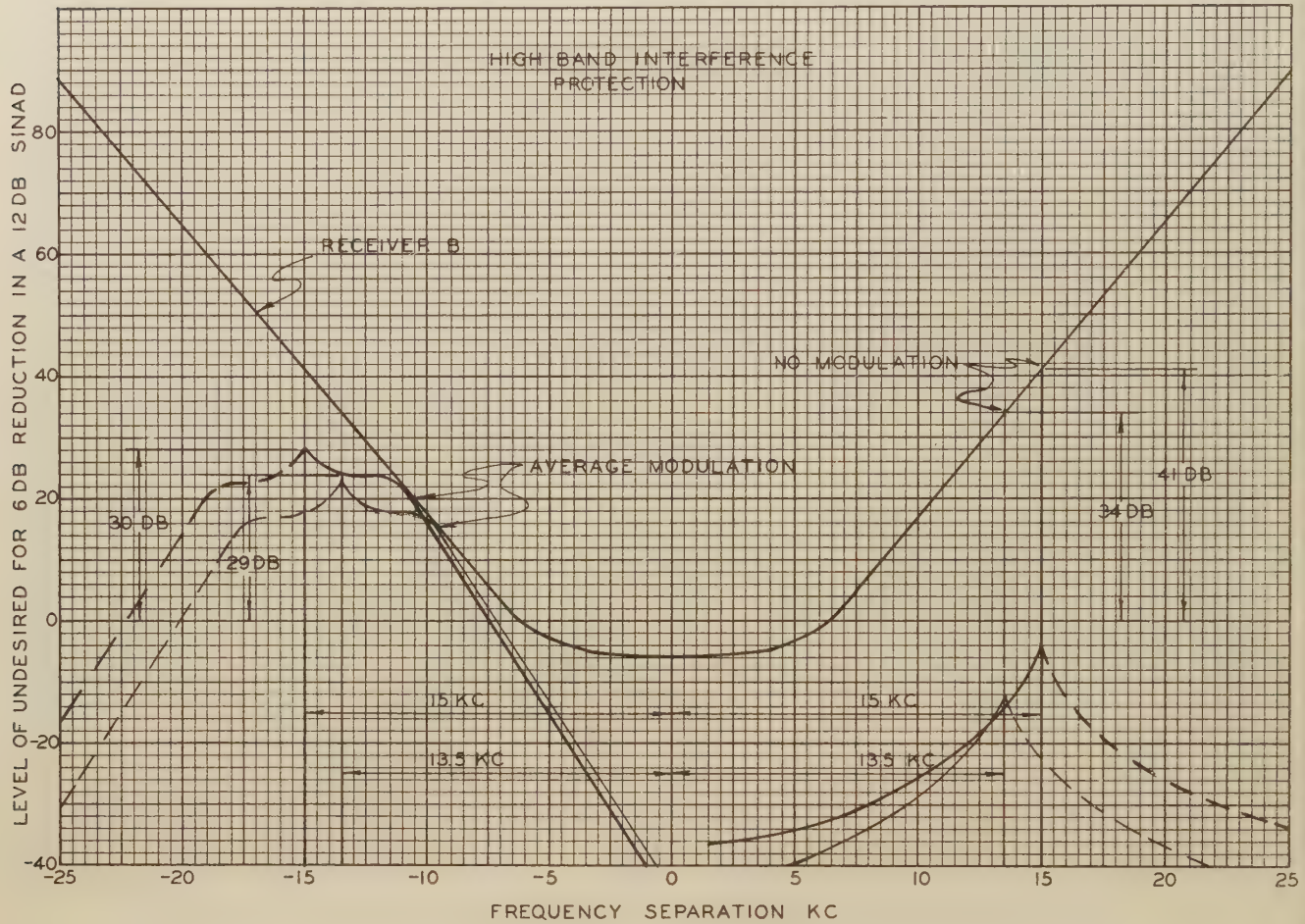


FIGURE 4
High Band Interference Protection

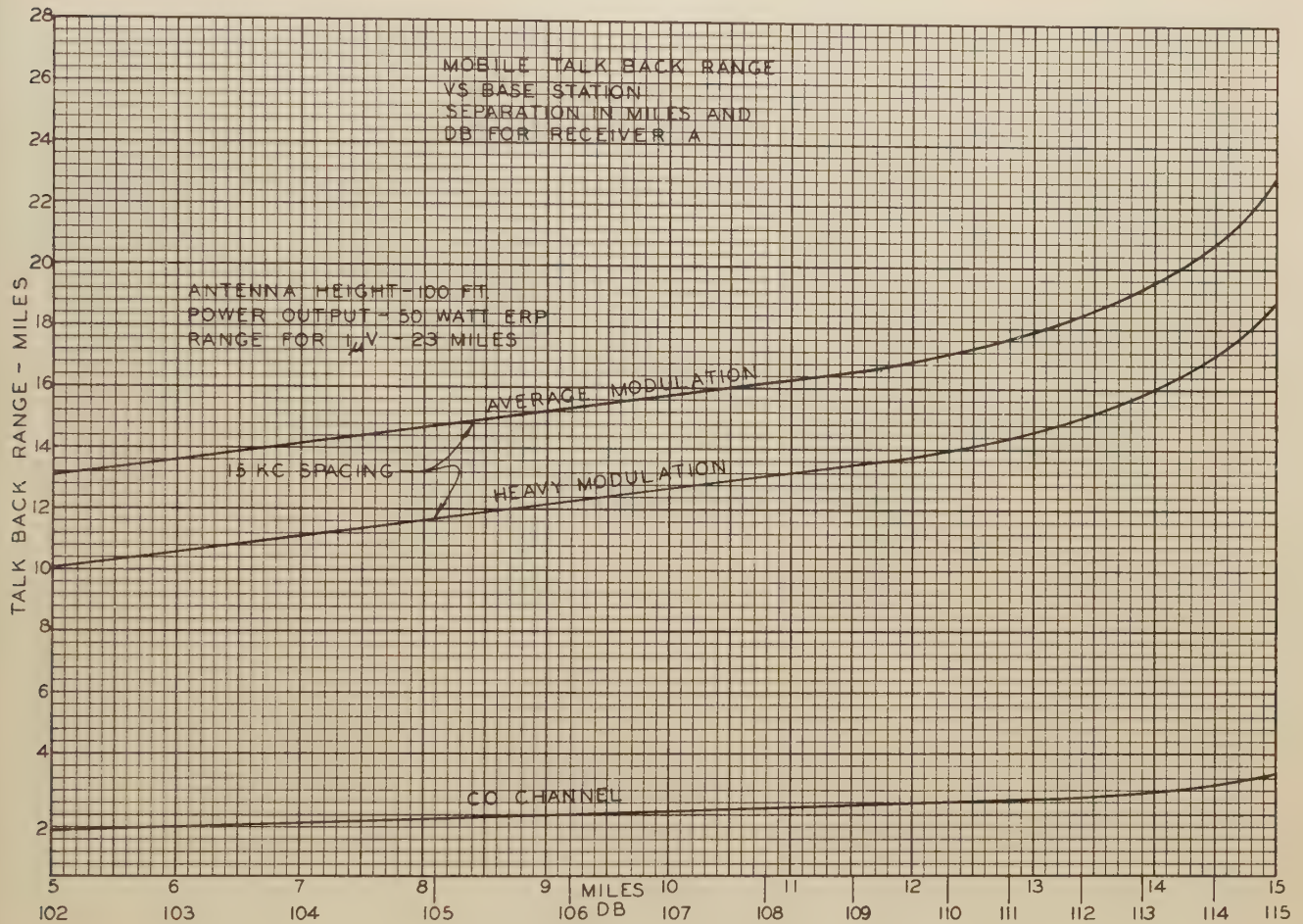


FIGURE 5

Mobile Talk Back Range vs Base
Station Separation in Miles and
DB for Receiver A

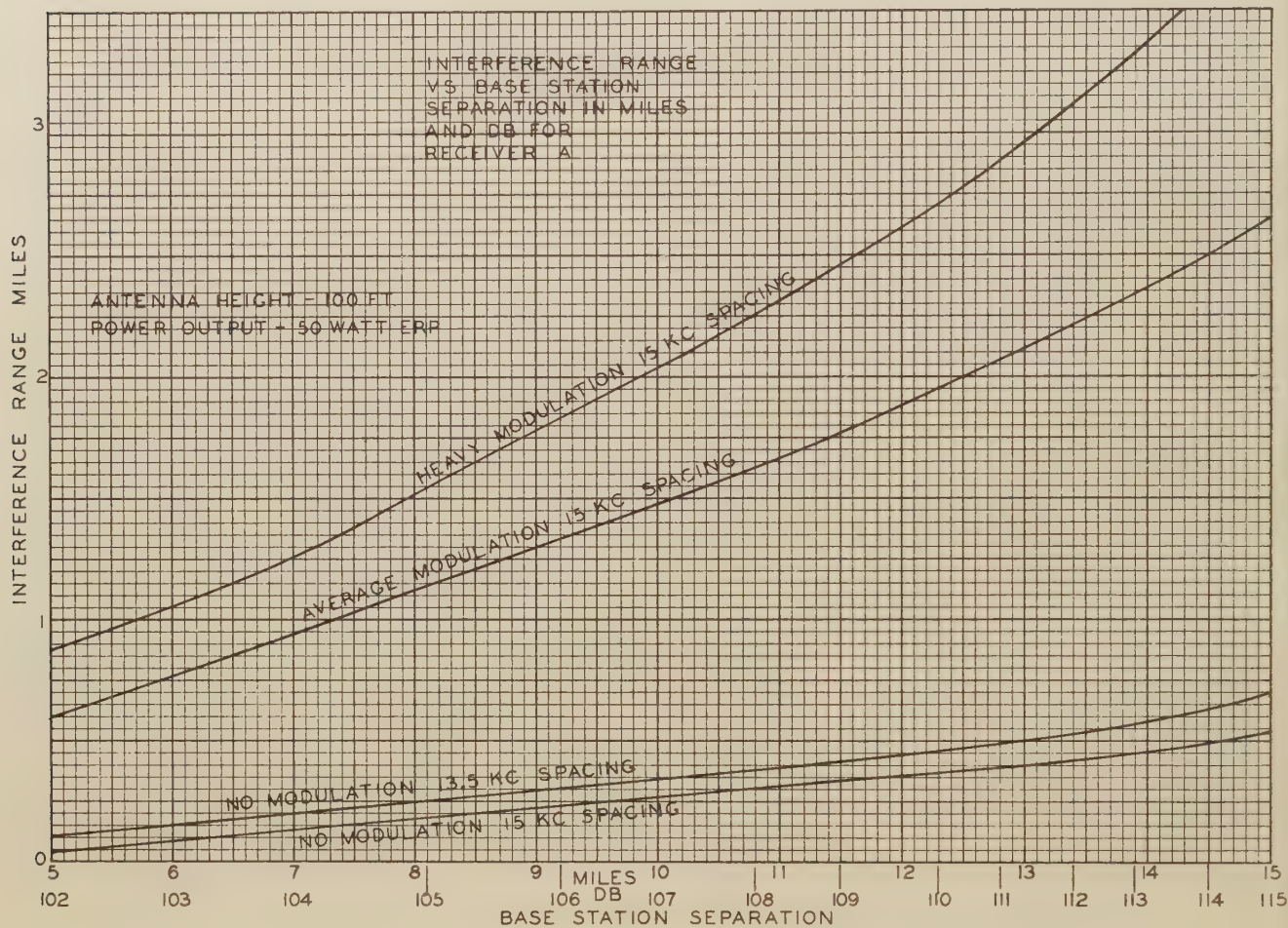


FIGURE 6

Interference Range vs Base
Station Separation in Miles
and DB for Receiver A

INTERFERENCE RANGES—(6DB DEGRADATION OF 12DB SINAD)

NARROW BAND—50 WATT TRANSMITTERS 100 FOOT ANTENNA

	BASE TO BASE		BASE TO MOBILE	
	High Band	Low Band	High Band	Low Band
CO-CHANNEL	56 MILES	62 MILES	29 MILES	40 MILES
15 KC SPLIT	15 MILES		3.5 MILES	
30 KC SPLIT	0.2 MILES		0.2 MILES	
10 KC SPLIT		45 MILES		28 MILES
20 KC SPLIT		3 MILES		1½ MILES

Figure 7
Interference Ranges

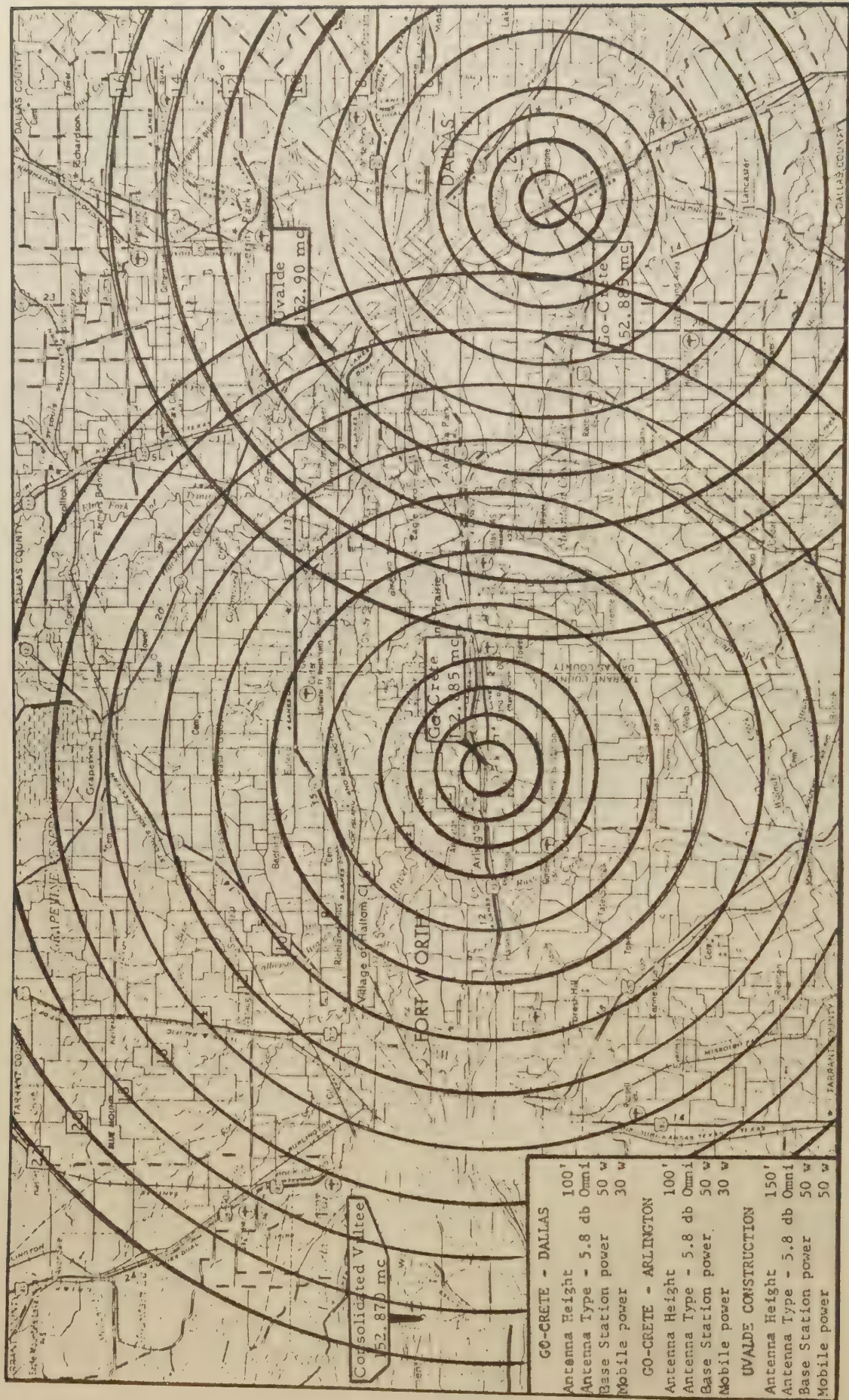


FIGURE 8

Map - Go-Crete and Uvalde Radio Systems

